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SUMMARY TECHNICAL REPORT
OF THE
NATIONAL DEFENSE RESEARCH COMMITTEE

[redacted]

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SUMMARY TECHNICAL REPORT OF DIVISION 13, NDRC

VOLUME 1

DIRECTION FINDER AND ANTENNA RESEARCH

OFFICE OF SCIENTIFIC RESEARCH AND DEVELOPMENT
VANNEVAR BUSH, DIRECTOR

NATIONAL DEFENSE RESEARCH COMMITTEE
JAMES B. CONANT, CHAIRMAN

DIVISION 13
HARADEN PRATT, CHIEF

WASHINGTON, D. C., 1946

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NOTES ON THE ORGANIZATION OF NDRC

The duties of the National Defense Research Committee were (1) to recommend to the Director of OSRD suitable projects and research programs on the instrumentalities of warfare, together with contract facilities for carrying out these projects and programs, and (2) to administer the technical and scientific work of the contracts. More specifically, NDRC functioned by initiating research projects on requests from the Army or the Navy, or on requests from an allied government transmitted through the Liaison Office of OSRD, or on its own considered initiative as a result of the experience of its members. Proposals prepared by the Division, Panel, or Committee for research contracts for performance of the work involved in such projects were first reviewed by NDRC, and if approved, recommended to the Director of OSRD. Upon approval of a proposal by the Director, a contract permitting maximum flexibility of scientific effort was arranged. The business aspects of the contract, including such matters as materials, clearances, vouchers, patents, priorities, legal matters, and administration of patent matters were handled by the Executive Secretary of OSRD.

Originally NDRC administered its work through five divisions, each headed by one of the NDRC members. These were:

Division A—Armor and Ordnance
Division B—Bombs, Fuels, Gases, & Chemical Problems
Division C—Communication and Transportation
Division D—Detection, Controls, and Instruments
Division E—Patents and Inventions

In a reorganization in the fall of 1942, twenty-three administrative divisions, panels, or committees were created, each with a chief selected on the basis of his outstanding work in the particular field. The NDRC members then became a reviewing and advisory group to the Director of OSRD. The final organization was as follows:

Division 1—Ballistic Research
Division 2—Effects of Impact and Explosion
Division 3—Rocket Ordnance
Division 4—Ordnance Accessories
Division 5—New Missiles
Division 6—Sub-Surface Warfare
Division 7—Fire Control
Division 8—Explosives
Division 9—Chemistry
Division 10—Absorbents and Aerosols
Division 11—Chemical Engineering
Division 12—Transportation
Division 13—Electrical Communication
Division 14—Radar
Division 15—Radio Coordination
Division 16—Optics and Camouflage
Division 17—Physics
Division 18—War Metallurgy
Division 19—Miscellaneous
Applied Mathematics Panel
Applied Psychology Panel
Committee on Propagation
Tropical Deterioration Administrative Committee

NDRC FOREWORD

As events of the years preceding 1940 revealed more and more clearly the seriousness of the world situation, many scientists in this country came to realize the need of organizing scientific research for service in a national emergency. Recommendations which they made to the White House were given careful and sympathetic attention, and as a result the National Defense Research Committee [NDRC] was formed by Executive Order of the President in the summer of 1940. The members of NDRC, appointed by the President, were instructed to supplement the work of the Army and the Navy in the development of the instrumentalities of war. A year later, upon the establishment of the Office of Scientific Research and Development [OSRD], NDRC became one of its units.

The Summary Technical Report of NDRC is a conscientious effort on the part of NDRC to summarize and evaluate its work and to present it in a useful and permanent form. It comprises some seventy volumes broken into groups corresponding to the NDRC Divisions, Panels, and Committees.

The Summary Technical Report of each Division, Panel, or Committee is an integral survey of the work of that group. The first volume of each group's report contains a summary of the report, stating the problems presented and the philosophy of attacking them, and summarizing the results of the research, development, and training activities undertaken. Some volumes may be "state of the art" treatises covering subjects to which various research groups have contributed information. Others may contain descriptions of devices developed in the laboratories. A master index of all these divisional, panel, and committee reports which together constitute the Summary Technical Report of NDRC is contained in a separate volume, which also includes the index of a microfilm record of pertinent technical laboratory reports and reference material.

Some of the NDRC-sponsored researches which had been declassified by the end of 1945 were of sufficient popular interest that it was found desirable to report them in the form of monographs, such as the series on radar by Division 14 and the monograph on sampling inspection by the Applied Mathematics Panel. Since the material treated in them is not duplicated in the Summary Technical Report of NDRC, the monographs are an important part of the story of these aspects of NDRC research.

In contrast to the information on radar, which is of widespread interest and much of which is released to the public, the research on subsurface warfare is largely classified and is of general interest to a more restricted group. As a consequence, the report of Division 6 is found almost entirely in its Summary Technical Report, which runs to over twenty volumes. The extent of the work of a Division cannot therefore be judged solely by the number of volumes devoted to it in the Summary Technical Report of NDRC; account must be taken of the monographs and available reports published elsewhere.

Of all the NDRC Divisions, few were larger or charged with more diverse responsibilities than Division 13. Under the urgent pressure of wartime requirements, the staff of the Division developed navigation and communications devices and systems which not only contributed to the successful Allied war effort, but will continue to be of value in time of peace in the fields of transportation and communications. The work of the Division, under the direction first of C. B. Jolliffe and later of Haraden Pratt, furnishes a foundation for what promises to be even more radical developments than those of the war—for one example, direction finders which will operate at all elevations and azimuth angles, in other words, hemispherically.

The Summary Technical Report of Division 13 was prepared under the direction of the Division Chief and authorized by him for publication. The report presents the methods and results of the widely varied research and development program, and, in the case of work with speech scrambling and decoding, it presents for the first time a comprehensive review of the state of the art. The report is also a notable record of the skill and integrity of the scientists and engineers, who, with the cooperation of the Army and Navy and Division contractors, contributed brilliantly to the defense of the nation. To all of these we express our sincere appreciation.

VANNEVAR BUSH, Director
Office of Scientific Research and Development

J. B. CONANT, Chairman
National Defense Research Committee

FOREWORD

NEARLY SIXTY years ago Heinrich Hertz experimentally produced electromagnetic waves, determined the direction of the waves, and wrote, "Thus we now have a means of discerning the direction of the electric force at every point." The waves were not detected outside of his lecture room, and it is unlikely that he foresaw the application of direction finding to navigation. Later, as the direction finder art advanced, many types of directional antennas were devised, including loops, crossed loops, spaced loops, Adcocks, and arrays. Some who contributed most effectively were Adcock, Ballantine, Barfield, Bellini, Busignies, Dellinger, Dieckmann, Eckersley, Hell, Kolster, Marconi, Mesny, Pickard, Smith-Rose, and Tossi.

During the fifteen years prior to World War II, the art advanced relatively slowly. Most progress was made in England. Equipment performance was reasonably satisfactory. Ground installations of direction finders were used to inform ships at sea of their positions. A similar use of ground direction finders was made by Pan American Airways and by various European air lines. Direction finders on ships at sea were almost universally used as a navigational aid, and most commercial airliners employed automatic direction finders for navigation. Thus, by the advent of World War II, direction finding was established as an important means of navigation.

Early in World War II, the Communications Division (Division 13) of the National Defense Research Committee [NDRC] formed a Direction Finder Committee under the Chairmanship of Loren F. Jones of which the members were H. Busignies, J. H. Dellinger, D. G. C. Luck, and R. K. Potter. Later, as consultants or technical aides, the Committee was greatly assisted by J. Allison, E. D. Blodget, and W. C. Lent. This Committee was active until September 21, 1945, with a number of Army, Navy, and British liaison representatives attending each meeting. During this period, the Committee issued contracts for work at Stanford University, California Institute of Technology, Harvard University, University of New Mexico, Federal Telephone and Radio Laboratories, Radio Corporation of America, Wilmotte Laboratories, J. A. Maurer, Inc., and Bell Telephone Laboratories. In addition, the Committee served as a coordinating agent and a clearing house for direction finder developments everywhere. The art advanced rapidly. Such diverse subjects as polarization errors, ionospheric effects, site errors, navigational applications, evaluation of fixes, and location of electric storms were studied.

Despite its long history, direction finding has been the subject of remarkably few texts. For years, the standard text in English was *Wireless Direction Finding* by R. Keen, published in England in 1922 and now undergoing its fourth revision. *Radio Direction Finders* by D. S. Bond was published in 1944.

The present publication, for which Keith Henney has acted as general editor and has devoted much time to coordinating the material, is Volume 1 of four books covering the wartime work of the Communications Division of NDRC. In this volume, there are accounts of developments sponsored by the Direction Finder Committee and of the results obtained. This book is not intended for the layman, and will be of only moderate assistance to equipment operators. It is intended for scientists, engineers, military personnel, students, and others who are interested in radio direction finding.

Radar, which combines direction finding and ranging, is already extensively used for navigation. To some extent, it will replace direction finding. However, direction finding will remain as one of the primary navigational methods and will be used for new functions such as locating electric storms. As the art advances, developments will facilitate direction finding at higher frequencies, will minimize errors, and will simplify equipment. Recent progress made in these directions by NDRC is outlined in the following pages.

The future holds promise of more radical developments, such as direction finders which will operate at all elevation and azimuth angles, in other words, hemispherically, with an accuracy adequate for fire control purposes. Possibly all directions and frequencies will be under continuous uninterrupted observation with some kind of panoramic presentation. Possibly there will be a need for direction finders with automatic tracking wherein the equipment will lock on and automatically follow a moving source of emission. No doubt there will be still other developments not now envisioned.

All radio communication, of course, involves the proper design and use of many components, among them antennas. Direction finding, radar, altimeters, and countermeasures for jamming enemy radio communication require means for imparting to and receiving from space the required radio energy. For this reason it was natural that certain research on the design, measurement, and application of antennas should fall to Division 13 to sponsor. Following the material on direction finding will be found summaries of the several antenna projects supervised by the Division.

HARADEN PRATT
Chief, Division 13

PREFACE

IN SUMMARIZING the several hundred reports of contractors on the hundred-odd research projects sponsored by Division 13 of the National Defense Research Committee, [NDRC], the editor has had to settle in his own mind how much or how little of each project report should be included; in other words, how far the boiling-down process should go.

The editor has an abhorrence for seeing good scientific or technical material go unpublished. Only by publication can the facts or methods developed by a few researchers become available for all researchers. On this basis, substantially all Division 13's program should be included in the volumes, of which this is one, summarizing the work of the Division. On the other hand, time moves forward inexorably so that it is quite likely that, by the day of publication, much of the data would already be out of date. Furthermore, time and human energy are always scarce. On these bases, all that might be required would be a paragraph or two summarizing the aims of the project and its accomplishments.

A middle course was steered, a course between the easiest solution of publishing practically all of each report and the more difficult job of really digesting the project purpose and results. The editor, however, deliberately chose to publish too much rather than too little. In most cases it will be unnecessary for the reader to search out the original source material unless he wishes to dig deep into the subject. In those cases where fundamental information was assembled and printed in the project report, that is, information on which future research might be based, the summaries have been permitted to take as much space as required.

This volume covers two aspects of Division 13's work—that dealing with research and development in direction finding, and that on antennas. The work on direction finders has been divided broadly into two aspects, that describing physical equipment, and that covering fundamental research leading to better knowledge of the manner in which ground constants, multiple rays, polarization by the ionosphere, and other factors affect the accuracy with which bearings can be measured. All this was necessitated by the fact that direction finding had gone into a sort of intellectual slump by the beginning of World War II. Antennas were generally of the loop or the Adcock type. Errors in bearings were deplored but accepted. Need had not risen for direction finding on the higher frequencies which came into such wide use during World War II. Above all, new ideas, new and basic analytical research were needed.

Throughout all the fundamental work on direction finding, the subject of errors was most important, simply because direction finders of various types do not give consistent nor accurate bearings in spite of the fact they can be erected with great care and constructed of precision apparatus. In fact, exploration of the vagaries of direction finding occupied a great deal of the attention of the Division and its research men and engineers. Finally, through the means of a new instrument, the polariscope, it was proved that many d-f troubles are due, not to the apparatus itself, nor to the ground on which it is located, nor to the operation of the equipment, nor to the fact that the ionosphere polarized radio waves heterogenously. Many of the errors which would remain, even if all the other sources of difficulty were removed, come from the fact that radiation from a transmitter arrives at a receiving point over multiple paths, and it is the many possible interrelations between these multiple rays that produce direction-finding aberrations. Thus it appears that there is a point beyond which much greater accuracy in bearing determination cannot be obtained by refining the apparatus.

Fundamental studies, analytical in nature, are reported rather fully in this report. Part I, dealing with basic studies in direction finding, includes means of measuring ground constants, and of rating d-f systems in terms of wanted-to-unwanted pickups; the effects of connecting cables with Adcock systems; a new means of controlling the amplification of a d-f receiver by means of a local transmitter; and the virtues of direction finding on pulse transmissions.

Part II deals with physical equipment and systems developed under the aegis of Division

13. Here will be found the work which led to the SCR-291, a single-band d-f system widely used by the Air Transport Service, a workable Radio-Sonde, a d-f system for the region of 140 to 600 mc, portable beacons which would lead a foot soldier to his objective on the field of battle regardless of weather or time of day, and means for locating tanks by radio. Finally, one of the last and most elegant accomplishments of the Division was an electrical and electronic instrument for evaluating the responses obtained from multiple d-f receivers so that the origin of signals could be more closely pinned down to a circle of small radius.

Part III records early work of sferics, the use of radio direction finding for locating storms.

The portion of the Division's activities dealing with antenna research is found in Part IV. Here is described the early work on determination of the characteristics of antennas for aircraft and tanks by means of scaled-down models; the work on faired-in antennas; a complete survey of airborne antennas as of early 1945, including what was then known about wide-band antennas. Work on disguised antennas, on improvised d-f antennas for use in the field, and on antennas for use in the region of 160 to 600 mc are also recorded here.

Keith Henney
Editor

CONTENTS

PART I

STUDIES OF HIGH-FREQUENCY DIRECTION FINDING

During the years immediately preceding World War II only a limited amount of basic research was devoted to the development of direction-finding [d-f] techniques and equipment. As in most other branches of scientific and engineering endeavor, the advent of the war accelerated such research first by making only too evident the need for it as relative to ordinary peacetime applications, and second by bringing into important focus new uses for d-f equipment. Much information was required on the use of d-f technique for use on high radio frequencies, on the causes of and solutions for certain vagaries in high-frequency direction finding, on the correlations between d-f measurements and the state of the ionosphere which reflects back to earth radio frequencies most likely to be used during the war.

Chapter 1

BTL HIGH-FREQUENCY DIRECTION-FINDER RESEARCH

Research leading to the general design requirements for circular-array direction-finder systems and Adcock antennas, including a determination of antenna spacing to minimize interaction effects, requirements for buried-conductor arrangements in both systems, design of antenna elements and coupling units to extend the frequency range of operation, specifications for receiver for a crossed buried U antenna direction-finder system, setting up of a complete direction-finder system using commercial receivers, and development of a unique method of automatic gain control. The text herein is condensed from the contractor's final report.[1]

1.1 STATE OF THE ART

At the time of this project,[1] the most promising high-frequency direction finders from the standpoint of simplicity and ease of operation were based upon the Adcock principle, but all such systems were subject at times to serious errors, the main cause being unwanted horizontal pickup in the antenna system. The most accurate high-frequency direction finders were the large fixed installations, but they were usually complicated, clumsy, and slow in operation.

The object of this project was to make a brief survey of the various types of high-frequency direction finders, to pick the most promising, to study the causes of the errors, and to determine, whenever possible, methods for reducing these errors.

The conclusion was reached that, for fixed installations and where speed of operation was important, Adcock systems were most promising. Accordingly, a crossed Adcock antenna system was designed and built. The errors in its receiving characteristics were studied, and methods were derived for their reduction. The final result was an Adcock antenna system with greatly reduced polarization errors. A receiving system was designed for operation with the antenna.

[1] Project C-16, Contract No. NDCrc-155, Western Electric Company.

1.2 INTRODUCTION

Fundamentally, operation of all radio direction finders depends upon the fact that the relative phases of the currents induced by a radio wave in two or more fixed, spaced wires vary as the direction of arrival of the wave varies. In some systems this phase difference is measured directly and the direction of arrival determined by a comparison of the measured phase difference with a previous calibration of phase difference versus direction. For convenience we will call this the *phase-comparison* method. The Navy's CXK direction finder, for example, works by this method. In other systems, instead of actually measuring the relative phases of the currents in the various antennas, the latter are connected together in such a way that, as a result of phase interference, different outputs are obtained from the antenna system for different directions of arrival. This will be called the *amplitude-comparison* method. Direction finders which use the loop antenna, the Adcock or any of its variations, or a sharp directional array are all examples of systems of this type.

For either case the determination of the direction of arrival from the amplitude or phase difference is straightforward when the signal arrives over a single path. Long-distance short-wave radio transmission, however, usually takes place by several paths of continually varying lengths. Due to the interference among the waves arriving over these various paths, in general the field strengths will not be identical at two or more spaced antennas and the phase differences will not be the same as for any of the component waves. However, if the directions of arrival are very nearly the same, the field strengths at the various antennas will also be very nearly the same and the phase differences very nearly what would have been obtained for a single wave arriving in the mean direction,

except for those periods when the relative phases of the component waves are such as to produce a weak signal at any place within the area occupied by the antennas. At such times the relative field strengths at the various antennas may differ greatly and the phase differences may differ by as much as $\pm 180°$ from the correct value. For these reasons accurate bearings cannot be obtained during the minima of a fading signal.

The closer the antenna spacing the shorter will be the period when the relative field strengths will differ appreciably and when the phase differences will be incorrect. When the directions of arrival of the various waves are radically different, the field strength and phase difference will vary rapidly and considerably, so that, in general, direction finding with simple antennas consisting of only a few elements becomes impossible.[b] However, if the directions of arrival are confined to a single vertical plane, the azimuth of the direction of arrival may still be measured with certain types of direction finders such as the crossed Adcock, described later, although the periods of weak fields when correct bearings cannot be obtained will occur very frequently.

1.3 PHASE-COMPARISON METHOD

A simple form of direction finder using the phase-comparison method would be one consisting of two fixed vertical antennas connected, through appropriate receiving and amplifying equipment, to a phase-measuring device. One way of accomplishing this phase measurement is to introduce a phase shifter, either in the radio- or lower-frequency branches of one of the receivers. This phase shifter is then varied until the outputs of the two receivers cancel (differ by 180°). The phase difference between the currents in the antennas is then found by subtracting 180° from the phase-shifter reading.

In Figure 1 let A and B represent two such antennas and let d be the distance between them. If a radio wave arrives at an angle β with respect to the line AB, then the phase difference ϕ between the currents induced in the two antennas will be given by the equation

$$\phi \text{ (degrees)} = \frac{360d}{\lambda} \cos \beta \qquad (1)$$

where

$$\cos \beta = \cos \alpha \cos \delta \qquad (2)$$

where α and δ are the horizontal and vertical angles of arrival respectively and λ is the wavelength.

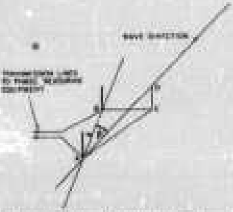


FIGURE 1. Diagram of simple phase-comparison direction finder.

Since β is measured from the array axis, equation (1) represents a cone whose axis is the line joining the two antennas and whose generator is at an angle β with respect to this axis. Thus all wave directions which lie in this cone will produce currents of the same phase difference in the two antennas, and, in general, additional information is needed to obtain the azimuth of the direction of arrival or the apparent bearing of the station.

This information can be obtained from another pair of antennas having a different orientation. The measurements obtained with this second pair of antennas will determine another cone with a different axis than the first. The line of intersection of these two cones will coincide with the actual direction of arrival of the wave and will, accordingly, determine not only the azimuth of the direction of arrival, but the vertical angle of arrival as well.

Disregarding, for the moment, the difficulties associated with the taking of two sets of data

[b] Highly directive steerable antenna systems such as the Musa type are required for this type of transmission.

simultaneously, a satisfactory direction finder might be made using two pairs of antennas arranged in two lines mutually perpendicular. In fact three antennas arranged at three corners of a square would answer the purpose.

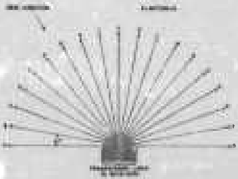
On the other hand, if two vertical antennas are mounted on a structure which can be rotated about a vertical axis until the phase difference between the currents in the two antennas is zero, then, if d is shorter than λ the apparent bearing of the station will be perpendicular to the line joining the antennas. Such a system cannot distinguish between signals having bearings 180° apart. To remove this 180° ambiguity requires the addition of a third antenna and greatly complicates the receiving equipment. This is the principle of the Navy's three-antenna CXK direction finder. One of the objections to this direction finder is the size of the rotating structure and the resulting time consumed in taking a bearing.

Circular Array

A variation of the foregoing scheme which overcomes the disadvantage just mentioned, at the expense of a slight decrease in accuracy, would make use of several fixed antennas spaced on the perimeter of a semicircle. These antennas would be used in pairs, any two adjacent antennas constituting such a pair. For making a measurement, that pair would be selected which was most nearly perpendicular to the direction of arrival and which, therefore, would give the smallest phase difference.

Figure 2 shows such an arrangement consisting of 19 antennas, one pair for every 10°. A wave is shown arriving at a bearing of 57° for which pair FG would be used to obtain the bearing.

The information obtained from a single pair of antennas is not, in general, sufficient to determine the apparent bearing of a station. However, if each pair is used to take bearings over only a small angular range approximately perpendicular to the line joining the antennas, then the phase difference of the currents in the two antennas can be used to determine the apparent bearing with a reasonable degree of accuracy for all but very high vertical angles of arrival, except for an approximate 180° ambiguity which could be removed only by the use of additional equipment. For a system such as is shown in Figure 2, where each pair is used over a range of only 5° on each side of the perpendicular, the maximum error for different vertical angles of arrival is given by curve A of Figure 3. Curve B gives the errors



[illegible]

for a system consisting of 10 antennas, one pair for every 20°. It will be observed that for the occasional signal suspected of having a high vertical angle of arrival the error may be eliminated by taking an additional measurement with another [illegible] tennas, preferably [illegible] pa[illegible] the axis of [illegible]ch is perpendicular to [illegible] of the first pair. If the phase-measuring device used for making this measurement is capable of operating over the full 360° range this measurement would also give the sense of the signal. In Figure 2, pair OP would be used to obtain this additional information.

As the separation between the antennas is increased, any given value of ϕ will correspond to smaller and smaller values of β [equation (1)]. Thus, for any given uncertainty in the value of ϕ, the greatest accuracy in bearing determination will be obtained with the largest possible value of d. However, for systems using only two sets of antennas with axes mutually perpendicular, the separation must be kept to

less than half the shortest wavelength upon which observations are to be taken. Otherwise, for some values of ϕ, there will be more than one possible value of β. For a system of several antennas located on a semicircle the separation may be increased somewhat as long as it is kept below $\lambda/2 \sin \theta$, where θ is the maximum angle on each side of the perpendicular at which

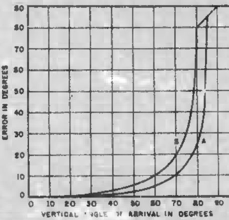


FIGURE 3. Errors in semicircle system. *A* shows errors when each antenna pair is used over range of only 5° on each side of the perpendicular; *B* shows similar errors for ten-antenna system, one pair for every 20°.

measurements will be made (5° for Figure 2). However, if this is done it will no longer be possible to make check measurements on the vertical angle of arrival with the in-line pair. For this reason it is recommended that the spacing always be kept below $\lambda/2$. Then, if greater accuracy is desired, after the approximate bearing is obtained a more widely spaced pair can be used to get a more accurate measurement. Thus in Figure 2, pair *NQ* could be used to get a more accurate measurement after a preliminary measurement is made with pair *OP*.

INTERACTION EFFECTS

The accuracy of all the above systems depends on the accuracy with which the phase difference between the currents in the separate antennas is given by equation (1). Among other things this phase difference is affected by the interaction among the various antennas that make up the system.

One of the objects of this project was the determination of the extent of this interaction and of the amount of error it would introduce in direction finders working by the phase-comparison method. In this study, use was made of some of the antennas of the broadside cage Musa system at the Holmdel, N. J., laboratories of the Bell System. These were vertical cage antennas $2\frac{1}{2}$ feet in diameter and $23\frac{1}{2}$ feet high. They had a half-wave resonant impedance of about 300 ohms at 18.15 meters and a quarter-wave impedance of about 36 ohms at 36.3 meters. The broad-band characteristic of this type of antenna makes it desirable for direction-finding systems which are to be used over a relatively wide frequency range. The low impedance makes it a simple matter to connect them to the receivers by means of low impedance, concentric transmission lines. The interaction will be a function of the dimensions of the antennas, but measurements were made with antennas of only one size since, in general, antennas used to cover the frequency range from 5 to 18 mc would be of approximately the same dimensions.

Figure 4 shows a ground plan of the antenna arrangement used for making these measurements. The antenna at point *X* was used as a reference antenna and all phases were measured with reference to the phase of the current in that antenna. Antennas 4, 5, and 6 in Figure 4 were antennas 4, 5, and 6 of the broadside Musa. They were fixed in location but could be easily lowered to the ground when not in use. These antennas as well as antenna *X* were all connected to buried coaxial transmission lines which terminated on the antenna termination panel in the Musa building and could, therefore, be connected to the Musa phase measuring equipment. Another exactly similar cage antenna was carried on a trolley suspended between the supporting poles of antennas 5 and 6. It was always connected through an 80-ohm terminating resistance to one of several ground rods which had been driven into the ground at approximately 6-foot intervals between anten-

nas 5 and 5. Antennas 4, 5, and 6 were spaced 49 feet (15 meters) apart.

By comparing the output of antenna 5 with that of antenna X while moving the traveling antenna between 5 and 5, with 4 and 6 lowered to the ground, the effect of an interacting antenna at distances of 6 feet to 49 feet was determined. By comparing the output of antenna 4 with that of antenna X while moving the traveling antenna between 5 and 5, with

FIGURE 4. Arrangement of antennas for interaction tests.

antennas 5 and 6 lowered to the ground, the effect of an interacting antenna at distances of 49 feet to 98 feet was determined. Measurements were made on five different wavelengths and with three different oscillator positions; one broadside to the antennas, at location A in Figure 4; one end-on, in the direction of the interacting antenna, at B; and one end-on in the opposite direction, at C.

In Figures 5, 6, and 7, the curves marked with open circles give the effect of the interaction on the amplitude of the current in the fixed antenna, those marked with crosses give the effect on the phase of the current, and those marked with solid circles give the corresponding error which the change in phase would introduce in the value obtained for β. For wavelengths between 16 and 64 meters a spacing of 49 feet does not introduce any significant error. This would be a perfectly satisfactory spacing for wavelengths greater than 30 meters but for shorter wavelengths this spacing becomes greater than $\lambda/2$ and would introduce an uncertainty in the bearings. For these shorter wavelengths a closer spacing is required, a spacing of 26 feet (8 meters) being satisfactory for wavelengths as short as 15 meters. The interaction for antennas at this spacing would introduce only a small error for

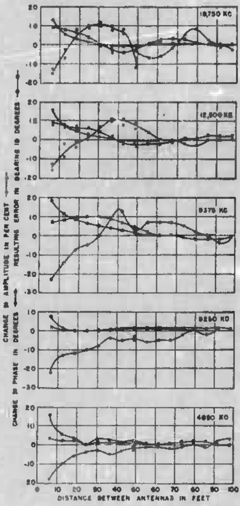


FIGURE 5. Interaction effects between antennas; oscillator at A in Figure 4.

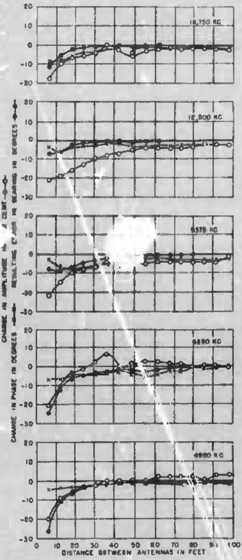


FIGURE 6. Interaction effects between antennas; oscillator at *B* in Figure 4.

wavelengths up to 24 meters but would be entirely unsatisfactory for wavelengths greater than 30 meters. Thus two complete antenna systems would be needed to cover the range from 16 to 64 meters (18.75 to 4.68 mc). It might be possible to balance out these interaction effects by using only the middle pair of

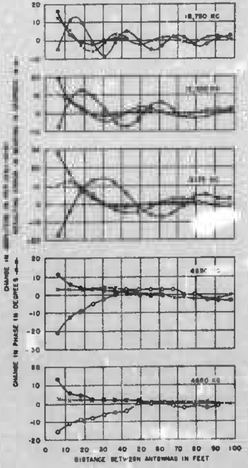


FIGURE 7. Interaction effects between antennas; oscillator at *C* in Figure 4.

a line of 4 or 6 equal-spaced antennas with the unused antennas terminated in a dummy load of the same impedance as the load impedance of the used antennas. However no tests have been made of such a system.

Phase-Comparison Method Possibilities

No attempt will be made to go into the details of the equipment that would be needed for a direction finder operating by the phase-comparison method. If only moderate accuracy is demanded a system could be built using several antennas arranged in a semicircle, each pair to be used for observing over only a limited range of azimuth. Once the correct antenna pair had been selected the taking of a bearing could be made practically instantaneous, but it might be necessary to try several different pairs before the correct one was selected and in that time the signal might be lost. It is conceivable that an instantaneous, direct-reading direction finder based on the phase-comparison principle could be devised, but the equipment would necessarily have to be very complicated and would require considerable time to develop. For these reasons attention was turned to systems working by the amplitude-comparison method.

1.4 AMPLITUDE-COMPARISON METHOD

Direction finders which use the balanced loop for a collector system are perhaps the most commonly known and simplest form of direction finder based upon the amplitude-comparison method. When properly constructed, they work very well for waves which are entirely vertically polarized; however, if there is any horizontally polarized component to the wave, currents will be induced in the loop which will mask the normal "figure eight" directional characteristic and will prevent the taking of accurate bearings. Since all radio waves which have suffered reflection from the ionosphere are more or less randomly polarized, this susceptibility to horizontally polarized waves makes the loop antenna practically useless for long-range direction finding on the short wavelengths.

Adcock Antenna

The Adcock antenna was designed to overcome the effect of horizontally polarized waves. In its simplest form an Adcock direction finder consists of two spaced vertical doublets connected by a balanced transmission line with a receiver connected across the transmission line at the midpoint. The conductors of the line to one of the doublets are reversed with respect to those to the other doublet. Connected in this way an Adcock antenna is, in reality, a two-element vertical array with the outputs in phase opposition. When the spacing between the doublets is small with respect to the wavelength the simple Adcock antenna has the same "figure eight" directional characteristic for vertically polarized waves as the loop. Figure 8A shows a schematic diagram of a simple Adcock antenna with its associated receiver and Figure 8B gives the horizontal directional characteristic.

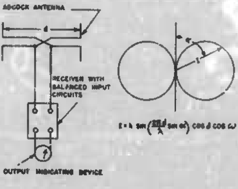


FIGURE 8. Diagram of simple Adcock direction finder.

In free space and with perfectly balanced transmission lines such a system would be unaffected by horizontally polarized waves, but, actually, the lower halves of the doublets are always nearer the earth than the upper halves so that a perfectly balanced system can not be obtained except through the use of compensating networks which require critical adjustment and must be retuned every time the receiver is tuned to a new wavelength. However, without these compensating networks, the unbalance is not serious if the antenna is elevated to a reasonable height above ground. Fairly accurate bearings can be obtained (1) if the rest of the equipment is kept small and simple, (2) if care is taken to keep the horizontal

members well balanced, and (3) if all the vertical elements other than the doublets are kept symmetrical.

Bearings are taken by rotating the antenna about a vertical axis until one of the nulls of the directional characteristic is pointed towards the direction of arrival of the signal, at which time the output of the receiver will at at a minimum. The direction of arrival or apparent bearing of the station will then be perpendicular to the line joining the antennas. The taking of bearings in this manner consumes an appreciable time, especially if the signal is weak or fading, when it is necessary to move the null of the directional characteristic back and forth across the signal direction several times before the bearing is certain. This time required for taking a bearing might mean that the signal would be lost before a bearing could be obtained. For this reason a quick-acting, direct-reading system would be desirable.

Direct-Reading System with Crossed Adcock Antennas

Figure 9A shows the schematic diagram of a direct-reading system using two mutually perpendicular Adcock antennas. In Figure 8B the equation for the output of a single Adcock was given as

$$I = k \sin\left(\frac{2\pi d}{\lambda} \sin \alpha\right) \cos \delta \cos \omega t \quad (3)$$

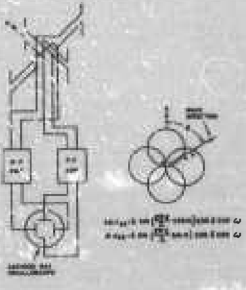
where I is the output current;
- k is a proportionality factor;
- α is the angle between the azimuth of the direction of arrival and the perpendicular to the line joining the two antennas;
- d is the spacing of the antennas;
- λ is the wavelength of the incoming signal;
- ω is 2π times the frequency of the incoming signal;
- δ is the vertical angle of arrival.

If now we have an antenna system consisting of two crossed Adcock antennas, and if we let the axis of one run north and south and that of the other run east and west, then the equations for the current output of the two antennas are

$$I_{NS} = k \sin\left(\frac{2\pi d}{\lambda} \cos \alpha\right) \cos \delta \cos \omega t \quad (4)$$

$$I_{EW} = k \sin\left(\frac{2\pi d}{\lambda} \sin \alpha\right) \cos \delta \cos \omega t \quad (5)$$

where α is now measured clockwise from the north-south line to the azimuth of the direction



A B

Figure 9. Diagram of direct-reading crossed Adcock direction finder.

of arrival and is, therefore, the apparent bearing of the station. If d is small with respect to λ the equations become[c]

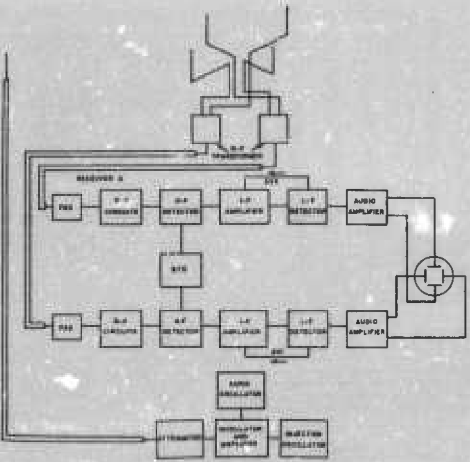
$$I_{NS} = k \frac{2\pi d}{\lambda} \cos \alpha \cos \delta \cos \omega t. \quad (4')$$

$$I_{EW} = k \frac{2\pi d}{\lambda} \sin \alpha \cos \delta \cos \omega t. \quad (5')$$

If these two antenna outputs are fed through appropriate equal-gain high-frequency amplifiers with equal phase shifts to the vertical and horizontal deflecting plates respectively of a

[c] When the spacing is not small with respect to a wavelength, i.e., when $\sin(2\pi d/\lambda)\cos\alpha$ is not approximately equal to $(2\pi d/\lambda)\cos\alpha$, an error is introduced which is zero for those bearings which are multiples of 45° and which reaches a maximum at the eight intermediate directions. Accordingly it is called the "octantal" error.

cathode-ray oscilloscope, the spot on the oscilloscope will trace a line which will make an angle α with respect to the vertical and will therefore, except for the 180° uncertainty, give an antenna system is small the interaction becomes large, but for the crossed Adcock antennas these effects are balanced out, providing the axes of the two pairs are exactly perpen-



[illegible] with injection signal.

the bearing directly. If the phase shifts through the amplifiers are not equal, the spot on the oscilloscope will, in general, trace an ellipse instead of a line, the major axis of which will not give an accurate bearing.

When the spacing between the elements of dicular and the antenna are all equispaced from the center.

It is to be noted that the achievement of this instantaneous, direct reading feature has required the complication of both the antenna system and of the receiving equipment, making

the problem of keeping the system balanced and symmetrical much more difficult. Some experimenters have attempted to overcome these difficulties by housing the receiving and indicating equipment in a shielded coop located at the exact center of the antenna system and then elevating the whole structure above the ground on poles. Even for such a system, care must be taken to keep the vertical portions of the power leads symmetrical and the horizontal portions well buried in the ground.

FIGURE 11. Details of construction of Adcock antenna system with square diagonals of 15 ft arranged north-south and east-west. Each antenna is 1½ ft in cross section, 28 ft high and is covered with copper-coated paper.

1.4 CROSSED BURIED U ANTENNA SYSTEM

If we had perfectly conducting ground another way of overcoming these difficulties would be to use only the upper halves of the doublets, bringing them down to the ground level and shielding the horizontal leads by burying them in the ground. It would be expected that such a system would work satisfactorily if the ground were uniform even if not perfectly conducting, and if the horizontal members were buried to such a depth that they were unaffected by the incoming waves. The paragraphs below contain a description of an antenna system of this type which was built at Holmdel and gives the results of various tests performed upon it.

Antenna Transformers and Buried Conductors

A schematic diagram of the antenna system and of the equipment used in testing it are shown in Figure 10. The four vertical antennas were located at the corners of a square with a diagonal spacing of 15 feet. The square was laid out so that one diagonal extended north and south and the other east and west. These four antennas consisted of box-like structures $1\frac{1}{2}$ feet square in cross section, extending 28 feet above the ground and covered with copper-coated paper. At 1 foot above the ground level

the four sides of the antenna were brought together at a point in the middle by an inverted pyramid of galvanized sheet iron while the four wooden corner posts were extended down 4 feet below the surface of the ground where they were bolted to a wooden framework.

Details of construction of these antennas are shown in Figure 11. Great care was taken when they were erected to keep them located

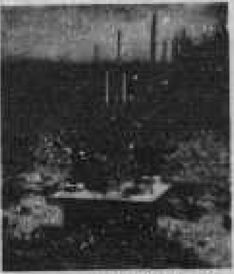
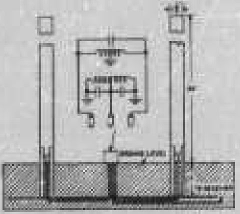


FIGURE 12. Conce[illegible] nected to broad-[illegible] shielded box at cen[illegible] is removed.

exactly as planned. After completion, check measurements showed a maximum error in direction of only 27 minutes and in spacing of less than ½ inch. The half-wave resonant impedance of a single unit (i.e., between the antenna and ground) was about 250 ohms and occurred at a wavelength of about 22.2 meters. The quarter-wave impedance was about 36 ohms and occurred at about 44.4 meters. Over the frequency range of 5 to 15 mc the output of a single pair for an end-on signal was about 9 db below that for a horizontal half-wave doublet at a height of 60 feet.

The inner conductor of a 72-ohm concentric transmission line was connected directly to the apex of each inverted pyramid, all four lines being of equal length. These transmission lines ran straight down to 4½ feet below the surface of the ground and then horizontally to the center of the system where they were brought back up to the ground level. Here they were connected to the primary, or balanced, side of a balanced-to-unbalanced broad-band transformer, the lines from each diagonal pair being connected to the same transformer. These transformers were housed in the small shielded box at the center of the antenna system which is visible in Figure 11. Figure 12 is a view of the interior with the cover removed.

Two more concentric lines, one from the secondary of each transformer, ran back down to 4½ feet below the ground, then horizontally for about 100 feet. Here they commenced a gradual rise to a depth of about 1 foot at which depth they remained until they reached the apparatus building at a distance of about 700 feet from the antennas where they were connected to the inputs of two receivers.



[illegible]

Figure 13 shows a diagonal cross section of the antenna system and illustrates the disposition of the transmission lines. They were buried in this manner to provide sufficient shielding to eliminate all pickup from the horizontally polarized waves.

Figure 14 shows the details of construction of the broad-band balanced-to-unbalanced transformers, and Figure 15 the frequency characteristics. Details of the measuring technique will be found in the project final report.[1] A balance of only 20 db would give an error in bearing of 5.7°. For a balance of 30 db the error is 1.8° and for 40 db the error is 0.6°.

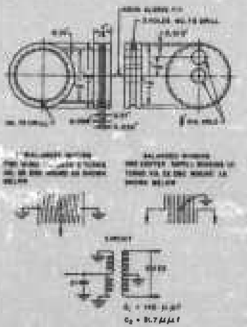


FIGURE 14. Construction details of balanced-to-unbalanced transformer.

RECEIVING ARRANGEMENTS—INJECTION-SIGNAL SYSTEM

Figure 10 is a block diagram of the d-f system. In operation, the incoming signal beats with the signal from the injection oscillator to give a beat note somewhere between 100 and 2,000 cycles. The injection-oscillator input level to the two receivers is equal at all times while the incoming signal input level varies in accordance with the directional pattern of the crossed Adcock antenna system. The receiving equipment measures the incoming signal direction by comparing the audio-frequency levels from the two receivers.

The audio output of the receiver connected to the north-south antenna pair is connected to the vertical deflecting plates of a cathode-ray oscilloscope and the output of the receiver connected to the east-west antenna pair is connected to the horizontal deflecting plates. Now it will be seen that if the overall gains of these two receivers are equal, the ratio of the audio outputs of the two receivers will be equal to the ratio of the carrier output of the two antenna pairs, which in turn is a function of the signal direction as shown above. When these voltages

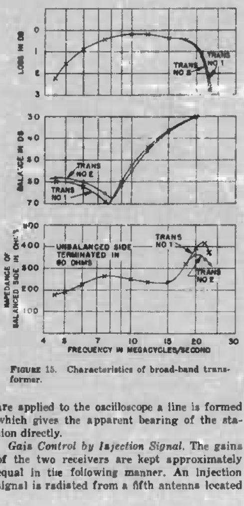


FIGURE 15. Characteristics of broad-band transformer.

are applied to the oscilloscope a line is formed which gives the apparent bearing of the station directly.

Gain Control by Injection Signal. The gains of the two receivers are kept approximately equal in the following manner. An injection signal is radiated from a fifth antenna located

at a distance of 500 feet from the crossed antenna system and on a line which bisects one of the 90° angles formed by them. This signal, which is adjusted to differ in frequency by a few hundred cycles from the carrier frequency of the signal whose direction is being measured, is used to control the gains of the two receivers. Since the outputs of both antenna pairs are equal for the injection signal the gains of the two receivers will be made equal providing the injection signal is much stronger than the carrier and providing the gain controls track. A ratio of injection signal to carrier signal of 26 db should be enough to insure that the carrier will have no effect on the gain controls and is enough to insure that the audio output produced in the linear low-frequency detectors by the beating of the carrier with the injection signal is directly proportional to the carrier output of the antenna system. Accordingly, the injection signal was kept about 26 db above the carrier level. The gain controls of the two receivers used for making preliminary measurements did not track within the required limits so that it was necessary to calibrate the system frequently. This was accomplished by modulating the injection oscillator with an audio frequency and adjusting the gains of the two receivers until the line on the cathode-ray oscilloscope was in the 45° position which is the condition for equal gains. This fault was eliminated in the final system described below.

Besides furnishing a nonfading equal-amplitude signal for controlling the gain of the receivers, the use of the injection oscillator greatly reduces the phase-shift requirements. Instead of equal phase shifts from antenna to oscilloscope, all that is required of the receivers when the injection oscillator is used, is that the phase shift from the antenna to the input of the audio amplifier varies in the same manner for the two receivers over the band between the carrier frequency and the injection oscillator frequency. The two audio amplifiers must have the same phase shift for the audio frequency used, but this requirement is not difficult to meet.

1.4 TESTS ON COMPLETE SYSTEM

First tests on the antenna system showed very shallow nulls and large errors in direction. For details of these measurements see the final report.[1]

When the possible causes for these errors were considered suspicion was first cast upon the vertical leads running to nearby antennas. Although the nearest of these antennas was over 200 feet distant, their removal and removal of their vertical leads made differences as great as 11 db in the depth and 3° in the direction of the nulls. Vertical wires at distances of 500 feet and over had no significant effect in the frequency range studied.

A set of measurements was made with all possible reradiating objects within a radius of 500 feet removed. Null depths of only 36 db and directional errors of 12° were still being obtained.

Effect of Ground

The ground at Holmdel is ordinary farm land consisting of a layer of top soil 1 foot thick overlying several feet of sandy clay.[2] The particular site chosen for the antenna system was as flat as reasonably could be expected of most antenna locations and, as far as could be discovered by visual examination, there was no reason for suspecting any troublesome variations in the ground constants. However, several different ground mats were tried. The first consisted merely of two 100-foot galvanized iron wires, one stretched under each pair of antennas and grounded at each end to 5-foot ground rods and at the center to the outer grounded conductor of the coaxial transmission lines. This ground system made no appreciable effect on either the depth or direction of the nulls.

Next a ground mat consisting of two 50-foot strips of 2-inch mesh wire netting 6 feet wide was laid in the form of a cross under the antennas. The maximum error in direction was reduced from 12° to 3.6°, although the minimum null depth was not changed much. Further improvement was desired, so a ground mat consisting of 8 radial strips of wire netting 6 feet wide and 150 feet long grounded at 50-foot intervals was tried. The results were not ap-

preciably different so still another mat was tried. This final one was a circular mat 100 feet in diameter grounded at 6-foot intervals around the circumference and at the center.

Except for the 18-mc measurements, the maximum null depth was over 30 db and the maximum error in direction was less than 2°. Since the antennas and transformers were not designed to work at frequencies higher than 15 mc it was felt that further improvement of the ground mat was not necessary.

With the 100-foot diameter ground mat in place, test bearings were taken on the field oscillator placed at 16 equally spaced points on a circle of 300 feet radius. When the errors were corrected for octantal error and for those due to the cathode-ray tube, over the frequency range 5 to 15 mc no error greater than 2° was observed, while at 18 mc the maximum error was 8°.

Tests on Horizontally Polarized Waves

To test the response of the system to horizontally polarized waves, a 53-foot tower was erected at a distance of 200 feet in the direction of the east null. The field oscillator was placed on top of the tower and the change in null depth noted when the antenna rods were turned from the vertical position to an angle of 45°. Measurements were taken on 5, 7.5, 10, 15, and 18 mc. No significant change in null depth was detected indicating that the amount of horizontal pickup was too small to affect the operation of the system under normal operating conditions.

Finally, bearing measurements were made on fixed transmitting stations ranging in frequency from 5 to 18.4 mc and in distance from about 30 miles to over 5,000 miles. Each bearing was checked by comparing it with that obtained on the same station with the crossed vertical Muss. A total of 107 bearings were taken. For 25 of these the vertical Muss gave either no bearing indication whatsoever or bearings differing significantly from the true bearing, indicating that either the station was inside the skip zone or that the transmission was unsatisfactory for direction-finder operation. Of the 82 remaining bearings, 29 differed from the true bearing by 0.5° or less, 24 differed by 0.6° to 1.0°, 12 by 1.1° to 1.5°, and 9 by 1.6° to 2.0° or a total of 74 of the 82 bearings were in error by 2° or less. For the remaining 8 stations the largest error was 6°. Carefully made repeat measurements on these 8 stations gave no error greater than 3°.

1.7 CONCLUSIONS

The errors in such a phase-comparison system, employing several vertical antennas arranged in a semicircle, are small unless the vertical angle of arrival is unusually high. Because of interaction effects, two complete sets of antennas are needed for the range from 5 to 18 mc if maxmium accuracy is required. With a system using three antennas, preferably arranged at three corners of a square, two simultaneous phase measurements are required to obtain a bearing. Interaction among the antennas influences the accuracy, although the interaction effects might be balanced out by the use of several dummy antennas.

Difficulties involved in making an instantaneous, direct-reading, *phase*-comparison system led to the development of an *amplitude*-comparison system using a crossed, buried U antenna with a separate injection oscillator and antenna and with a cathode-ray indicating device.

Variations in the ground not detected by the eye cause severe distortion of the directional pattern of the antenna. Several ground-mat systems were investigated. Conservative specifications indicate that a mat not less than 150 feet in diameter made of 1-inch mesh wire netting or its equivalent would be required. Where the ground has a uniformly high conductivity such as would be found in a salt marsh, the mat could be smaller. Even in a location having good ground conductivity, a good ground mat seems desirable. For details of a system suitable for a salt-marsh location see the final report.[1]

1.8 RECEIVER SPECIFICATIONS

The general receiver characteristics desired are (1) a frequency range of 4.5 to 30 mc, (2) an input impedance of 72 ohms, and (3) an image rejection ratio of better than 50 db at

20 mc. An l-f band flat over a ± 2-kc range and down 45 db ± 10 kc would be satisfactory. To operate ordinary commercial oscilloscopes an a-f output of 2.5 volts across 100,000 ohms at 5 per cent modulation is required.

The lowest signal level that can be received is determined by the equivalent input noise (output noise divided by the receiver gain) which should not be more than 5 db above the theoretical thermal noise or 160 db below 1 watt for a 4,000-cycle band width. Assuming a minimum signal level 20 db above the noise gives a minimum signal input level of 140 db below 1 watt.

The ratio of the a-f output levels must be the same as the ratio of the incoming signal levels delivered by the two antenna pairs. Therefore, if linear low-frequency detectors are used the level from the injection oscillator must be at least 20 db above the incoming signal level. The effect of the incoming signal on the automatic-gain-control circuits may require a 30-db difference in level. If we assume a 26-db difference in level, the modulation impressed on the injection oscillator signal by the incoming signal will never exceed 5 per cent. If incoming signal levels of 140 to 80 db below 1 watt are to be accommodated the injection oscillator level must be between 114 and 54 db below 1 watt and the receivers must be capable of handling these levels. The input level range may be increased to 80 db by inserting 20-db attenuation in the antenna lines to take care of exceptionally strong signals.

For successful operation, the gains of the two receivers must be kept equal at all times. This feature practically demands the use of independent and extremely "stiff" automatic gain controls whose action is completely controlled by the injection oscillator. These stiff gain controls may take the form of separate i-f gain-control amplifiers and strongly biased rectifiers.

For a bearing accuracy of $\frac{1}{3}°$, the gains of the two receivers must not differ by more than 0.1 db, and this must hold over the entire range of input levels. To eliminate any audio frequency gain controls so that the amplitude of the oscilloscope trace may be used as an indication of the percentage modulation (i.e., ratio of incoming signal to injection oscillator level) the automatic gain controls should keep the output constant to within 2 db for a 60 db change in input level.

Three other factors which will affect the bearing accuracy are (1) nonlinearity of the a-f circuits, (2) dissimilarities in the r-f, l-f, and a-f bands, and (3) crosstalk between the two receivers. To maintain an accuracy of $\frac{1}{3}°$ the a-f circuits should not depart from linearity by more than 0.1 db as the amplitude is varied, and the gains should be alike to within 0.1 db as the frequency difference between the injection oscillator and the signal is varied over a ± 200-cycle to 2-kc range. The crosstalk from one of the receivers, fed with a 5 per cent modulated signal, into the other with an equal unmodulated signal, should be down 50 db.

Factors which will distort the oscilloscope trace are 60-cycle hum, harmonic distortion, and phase shift. The hum-and-harmonic distortion should be down 35 db in the output in order that the oscilloscope trace will not be widened by more than 2 per cent of its length. The phase difference between the two a-f outputs should not exceed 1 per cent.

If necessary, high-pass filters may be used in the a-f circuits to reduce the 60-cycle hum, in which case the a-f range might be 200 to 2,000 cycles. Filters cutting off above 2,000 cycles might be used to reduce noise. Any such filters must, of course, meet the phase and amplitude-distortion requirements within the pass band.

The receivers, preferably, should have a frequency-calibrated dial and a tuning indicator to insure that the operating requirements will be met. Separate beating oscillators may be used, provided the leakage from one beating oscillator into the other receiver does not result in a-f components stronger than 50 db below the output of the desired frequency.

1.9 COMPLETE D-F SYSTEM

No commercial receivers were on the market which exactly met the specifications listed above, and previous experience with commercial short-wave receivers showed that their modification to meet these specifications would not be easy. Since, however, the construction of two entirely new receivers would have taken

a relatively long time, it was decided to use, after modification, two short-wave measuring sets belonging to the Bell Telephone Laboratories.

Receivers Used and Changes Required

A block diagram of one of the two receivers before alteration is shown in Figure 16. They are triple-detection receivers with a first intermediate frequency of 3 mc and a second intermediate frequency of 100 kc. There are no r-f amplifiers or preselector circuits other than the antenna-tuning and coupling circuits. The receivers cover a frequency range from 4.5 mc to well over 30 mc. The image rejection ratio is about 40 db which is high enough for demonstration purposes.

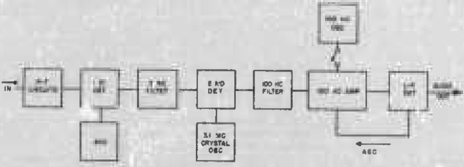


Figure 16. Block diagram of triple-detection measuring set.

The overall pass band of the receivers was flat for approximately 5 kc on each side of the carrier, which was entirely too wide for d-f use. By adjusting the second i-f filters the band was reduced so that it was flat for only 1 kc on each side of the carrier and was down 45 db or more 10 kc on each side of the carrier. It should be noted here that the response curves of the two receivers must be exactly similar over the operating range. This range is determined by the difference between the signal frequency and the injection-oscillator frequency. If the band width is small and the response curves sharply peaked it becomes extremely difficult to make and keep them identical.

The linear rectifiers used as final detectors were capable of supplying an audio output well over 2.5 volts across 100,000 ohms, but the overall gain of the receivers was not sufficient to supply this output for the weakest usable signal which is determined by the minimum noise level of the receiver. By adding another stage to the 100-kc amplifier the overall gain was increased to the point where the required audio output was obtained for the weakest usable signal.

The equivalent input noise was about 5 db above the theoretical thermal noise level.

It was necessary to install separate, biased, gain-control rectifiers to obtain the desired automatic-gain-control stiffness. The result was an output variation of only 1.6 db for a 60-db change in input level. The gain controls of the two receivers tracked to better than 0.1 db over the range of inputs from 65 db above to 25 db above 10^{-12} watt. At lower input levels the noise prevents precise measurement of the tracking, but no appreciable difference was discernible.

The hum level of the receivers was fairly high but was reduced to a satisfactory value by minor changes in the power supply leads. The nonlinearity in the audio circuits, the harmonic distortion, and the difference in phase shift were all less than the required minimum.

After the above changes were made the two receivers were connected together and to the antenna system making the complete d-f system shown in the block diagram in Figure 17. The first beating oscillator from one of the re-

ceivers was used as the common beating oscillator for both receivers. The first beating oscillator for the other receiver was removed and mounted on a separate panel with a broad-band amplifier and used as the injection oscillator. The injection-signal level was controlled by cycle tone and adjusting the relative audio gains of the two receivers until the oscilloscope trace made a line at 45°. The 60-cycle modulation was produced by putting a small 60-cycle voltage on the grid of the amplifier tube in series with the grid bias.

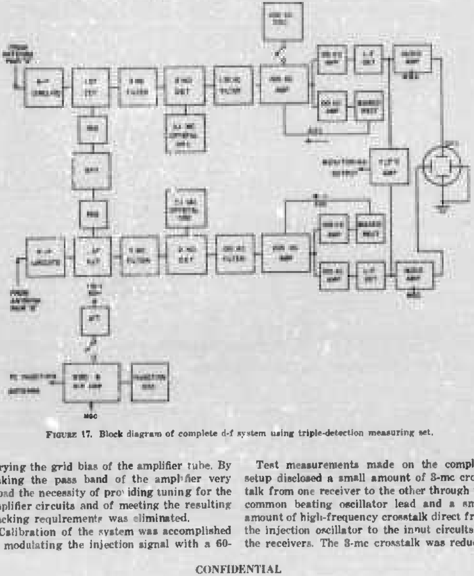


FIGURE 17. Block diagram of complete d-f system using triple-detection measuring set.

varying the grid bias of the amplifier tube. By making the pass band of the amplifier very broad the necessity of providing tuning for the amplifier circuits and of meeting the resulting tracking requirements was eliminated.

Calibration of the system was accomplished by modulating the injection signal with a 60-

Test measurements made on the complete setup disclosed a small amount of 3-mc crosstalk from one receiver to the other through the common beating oscillator lead and a small amount of high-frequency crosstalk direct from the injection oscillator to the input circuits of the receivers. The 3-mc crosstalk was reduced

below the required minimum by inserting small pads in the leads to each receiver and increasing the beating oscillator power by a corresponding amount. The high-frequency crosstalk was reduced by very carefully shielding the injection oscillator.

The experience gained in working with these receivers showed that the greatest difficulties in building receivers for this type of d-f system are likely to be in making the two pass bands identical over the operating range, in keeping the crosstalk through the common beating oscillator lead at a low value, and in keeping the leakage from the injection oscillator direct to the receiver below the required minimum. By taking normal precautions all requirements were met, demonstrating that the system is entirely practical.

1.10 TECHNICAL AID FOR THE ARMY SIGNAL CORPS

After a thorough testing, the two measuring sets and the necessary cathode-ray equipment were set up in a small building a short distance from the crossed buried U antenna system and connected to it by coaxial transmission lines making a complete d-f system. Engineers from the Signal Corps Laboratories then operated and studied this equipment to familiarize themselves with the principles involved.

Receivers Used and Changes Required

After operating this equipment for a while and after considering the possible sources of supply and the urgency of the need the Signal Corps engineers decided to attempt to rebuild two National N. C.-100 receivers for their first system. At first it was hoped that this rebuilding would involve merely the addition of a separate automatic-gain-control amplifier and rectifier and of a lead between the two beating oscillators to keep them in synchronism. However, before the equipment was finally made to function satisfactorily, it was found that rather extensive changes had to be made. In the final arrangement three receivers were used, one for each of the receiving channels and the third to supply the beating oscillator and injection signals.

Test Results. The final tests on these receivers showed that they functioned very satisfactorily except for two features. The image-rejection ratio was better than 40 db for frequencies between 5 and 11 mc, but above 11 mc the rejection ratio dropped very fast until at 14 mc it was only 28 db and even less at 15 mc. The equivalent input noise of these receivers was considerably higher than the specified minimum, especially on the higher frequencies, making it difficult to obtain accurate bearings on weak stations. It is suspected that the difficulty lies in the comparatively low gain of the r-f amplifier and in the low Q of the h-f coils.

It is not believed that low image-rejection ratio at the high frequencies is so serious as to rule out the use of such receivers, but, since it is the weak signals that are the important ones, the low signal-to-noise ratio is very serious.

1.11 EXTENDING THE RANGE TO 30 MC

It was felt that, as a result of the experience gained in building and testing the crossed buried U antenna system for the 5- to 15-mc range, enough was known about the characteristics of such an antenna system to predict the performance of a smaller model with sufficient accuracy to make unnecessary the building of an experimental model. The antenna system for the 15- to 30-mc range would be merely a half-size scale model of the present system. Thus, the ground mat should be 75 feet in diameter, the antenna spacing should be 7½ feet, and the size of each individual vertical should be 9 inches by 9 inches by 14 feet.

The Transformers

The design for the broad-band balanced-to-unbalanced antenna-coupling transformers followed very closely that of the ones for the lower frequency range. For details of construction of the transformers and loss, balance, and impedance characteristics see the final report.[1]

1.12 POLARIZATION ERRORS—SYSTEM MEASUREMENTS

D. G. C. Luck and Kenneth A. Norton of the RCA Manufacturing Company brought their balloon equipment to Holmdel and made measurements of the response of the crossed buried U-antenna system to vertically and horizontally polarized waves for various vertical angles of arrival. The standard wave error[a] was found to be 8.5° at 5.1 mc, 6.25° at 9.23 mc, 2.2° at 13 mc, and 1.8° at 17.3 mc.

Chapter 2

NBS HIGH-FREQUENCY DIRECTION-FINDER RESEARCH

Theoretical and experimental investigation of direction-finder characteristics, particularly polarization errors; development of a figure of merit for direction-finder comparison; examination of typical direction-finder systems as an application of the methods for measurement and analysis developed; origin of a new method for measuring ground constants. The major portion of the theoretical analysis developed in this project[a] is included in this summary report, the chief abridgement from the contractor's final report[1] being in the work conducted on direction-finder systems of the time 1941–1942.

2.1 INTRODUCTION

THE OBJECTIVES in setting up this project were:

1. Study of errors due to polarization, collector spacing, and diversity factor, and methods to minimize these.
2. Study of errors of site and personnel.
3. Examination of improved models from any source.
4. Basic research on one or more improved types as appears desirable.
5. Measurement technique for the study of d-f errors.

After some preliminary study it was found that polarization and site errors constituted the largest errors in existing direction finders. The program therefore was chiefly devoted to a study of those errors over a frequency range of 2 to 30 mc.

2.2 ANALYSIS

For the study of polarization errors a method was developed having advantages over previously used methods and applicable to many d-f antenna systems. In this method a figure of merit designated as the "pickup ratio" was introduced. The pickup ratio is the ratio of the pickup factor, h, of the d-f antenna system for desired radiation field components to its pickup factor, k, for undesired field components. A knowledge of the pickup ratio together with the directional pattern of response of the d-f system makes possible the determination of the polarization errors for downcoming sky waves. Since it is possible to measure the pickup ratio for a wave at horizontal incidence, all measurements may be made near the ground. This is a principal advantage of the method; other advantages are that the method yields the *maximum* polarization error, and a figure of merit for polarization error which is independent of the ground constants and of the height of the direction finder above the ground.

After developing the technique of determining polarization errors through measurements of pickup ratios, measurements were carried out on several direction finders of various types. Reports issued during the project are listed in the Bibliography.

Polarization errors were investigated comprehensively, both theoretically and experimentally. The polarization of the field at a d-f site for downcoming ionospheric waves was determined theoretically. The d-f directional pattern was then calculated in such a field and equations obtained for the observed bearing. The difference between the observed bearing given by the actual directional pattern and the true bearing obtained from the ideal directional pattern gave the polarization errors. Equations were derived for the polarization errors in this way for the several basic direction finders and were used to determine, by means of experimental measurements of the constants h and k, the polarization errors of these direction finders.

2.2.1 Summary of Theoretical Aspects

Study of the state of polarization of downcoming ionospheric waves showed that these waves were elliptically polarized, having electric components E_p and E_s polarized parallel and perpendicular respectively to the plane of

[a] Project C-18: The work covered in this report was performed by the National Bureau of Standards under a contract terminating June 30, 1942.

incidence. These components are present independently of the state of polarization of the wave incident on the ionosphere and therefore of the polarization of the transmitting antenna. The wave incident on the ionosphere is split into ordinary and extraordinary waves which on returning to the earth combine vectorially to give the total downcoming wave. Equations for the fading of these ordinary and extraordinary wave components were derived and expressions found for the *variations* in the state of polarization of the downcoming wave. In general, the state of polarization varies in a random way so that the average of a series of swinging bearings will usually give a bearing close to the true bearing, provided the swinging is caused by polarization error.

The *total* field at the direction finder for downcoming waves was next calculated by taking the vector sum of the direct and ground-reflected waves. It was shown that the ground reflection acts to suppress E_θ at points near the ground, the suppression increasing as the index of refraction of the ground increases. This showed that direction finders designed to respond to E_ϕ and to suppress response to E_θ should be placed as near the ground as possible. On the other hand, it is shown that a direction finder designed to respond to E_θ and to suppress response to E_ϕ should be located at a height above ground equal to $\lambda/4$.

The response of an arbitrary direction finder in the field of a downcoming wave was next calculated in terms of the known directional patterns of the antenna elements of the direction finders. In these expressions unknown proportionality constants, h and k, occur. These constants, which correspond to output voltages produced by the desired and undesired field components respectively, were called pickup factors, and the ratio of h to k the pickup ratio. It was shown that, by placing the direction finder in plane-wave fields of special structure, all terms in the expressions for the output voltages became zero except one. A measurement of the output voltage and the field intensity for this case provided a means for determining the pickup factor. The pickup factor for each field component desired could be measured in this way by using special fields at horizontal incidence. After determining the pickup factors experimentally, the d-f response in the field of any downcoming wave could be calculated and therefore also the polarization errors, since this calculation gave the actual azimuthal directional pattern. The departure of this directional pattern from the ideal desired pattern gave the bearing error.

It was shown that the polarization errors were dependent on the ratio of desired to undesired responses and therefore on the pickup ratio h/k. The pickup ratio was therefore proposed as a figure of merit for measuring polarization errors. This figure of merit is independent of the ground constants of the d-f site and of the height of the direction finder above the ground. It can be used to determine the results of development work designed to reduce the polarization errors of a given direction finder while the complete curve of polarization error versus angle of elevation of the incident wave, as determined by measurements of h/k, can be used to compare the accuracy of different types of direction finders.

In applying the pickup ratio method in practice, the required special test fields must be generated by means of local transmitters. Such transmitters generate waves which only approximately simulate the plane waves assumed in the theory used to calculate the polarization errors. Accordingly, theoretical and experimental studies were made of the techniques required for a proper measurement of h and k where a local transmitter is used. It was shown that a horizontal loop antenna should be used with the local transmitter when generating a horizontally polarized test field to avoid an error called *radiator* parallax. Similarly, when testing a spaced, vertical, coaxial loop-antenna direction finder, special procedures were necessary to avoid an error called *collector* parallax. Equations were also derived to show the proper procedure required when measuring electric field components by means of a field intensity meter using a loop antenna.

Using the procedures outlined above, measurements of h and k were made and the polarization errors computed for the direction finders under consideration.

A study of d-f sites was made in which it was shown that direction finders designed to respond to E_ϕ should have smaller site errors

caused by reradiation than those designed to respond to E_p. Equations were derived for the field intensity at any given depth below the ground for incident downcoming waves and an approximate table prepared showing the recommended depth to which cables and lines should be buried in order to avoid reradiation difficulties. A new method was evolved for rapidly measuring the ground constants of a proposed site, at various points of the site, both to determine its electrical homogeneity and to make sure that its conductivity and dielectric constant would be high enough for the best results with the direction finder to be installed. This method should be useful in selecting the best site when a choice is possible.

2.2.2 Historical Development

The single loop-antenna direction finder has large bearing errors when used on downcoming waves from the ionosphere. Since long-distance communication makes use of propagation via the ionosphere, the single loop-antenna direction finder could not be successfully used for short-wave direction finding in the band from 2 to 30 mc. In 1919, Adcock introduced the spaced-antenna direction finder which attempted to reduce pickup in the horizontal members of the antenna structure and thus the polarization errors. To measure the success of such attempts, R. H. Barfield[2] in 1935 introduced a figure of merit for polarization error which he called the "standard wave error." This was defined as the bearing error of the direction finder for an incident wave having an angle of elevation of 45° and components of electric field intensity parallel and perpendicular to the plane of incidence equal to each other and of such phase difference as to make the error a maximum. The standard wave error was commonly measured by using a local transmitter elevated at 45° to lay down a field simulating the standard wave. A dipole transmitting antenna oriented at 45° generated equal and cophased parallel and perpendicular wave components in this method. The cophased wave components of the above method did not result in the measurement of the maximum polarization error as required in the definition of standard wave error.[3] Furthermore, the difficulties introduced in the use of elevated transmitters led to lack of precision in measurement. A method was needed which would be easier to use in practice; for example, one which required ground measurements only. Such a method was developed using a local transmitter near the ground to generate a wave field of measured intensity at nearly horizontal incidence. The response of the direction finder was measured in this field for vertical and for horizontal polarization. The response of the antenna system to sky waves was then determined from a calculation of the vertical and horizontal field components of such sky waves, and this in turn gave the polarization errors, including the standard wave error. This method yielded the maximum polarization error, while the ratio of the responses or pickup factors, called the pickup ratio, yielded a fundamental constant of the antenna system which was independent of the ground constants and from which the response in any assumed sky-wave field could be calculated. About the same time as the development of this method, the Radio Corporation of America [RCA][4] modified the Barfield method by using an elevated transmitter emitting waves polarized, first, in the plane of incidence and, second, perpendicular to the plane of incidence. The response of the direction finder to these waves was measured, but a measurement of field intensity was not required to determine the polarization error. This method usually could be made to give the same data as the National Bureau of Standards [NBS] method and vice versa, each method having particular advantages.

2.2.3 Nature of Polarization Errors

In short-wave direction finding, bearings are taken on sky waves coming down from the ionosphere. In general these waves are elliptically polarized, having components polarized both parallel and perpendicular to the plane of incidence. However, most direction finders are designed to measure the bearing by utilizing the directional pattern of the antenna-system response to only one of the components. If both components are present and if the antenna system has different azimuthal responses to the several field components, the directional pat-

tern will be modified so as to give incorrect bearings or even to prevent the taking of bearings altogether. These bearing errors will depend on the relative amount and phase of desired to undesired field intensity which in turn will depend on the state of polarization of the incident wave. Such bearing errors are called polarization errors and are one of the largest sources of inaccuracy in present-day direction finders.

2.2.4 Polarization of Downcoming Ionospheric Radio Waves

As a background for the study of polarization errors and the techniques used to measure such errors, it will be desirable to consider the nature of the downcoming sky waves, the influence of the ground reflection on the fields set up by these waves at the direction finder, and finally the problems involved in attempting to simulate such sky waves by the use of local transmitters. A more complete account than can be given here is available in a report[1] by K. A. Norton.

The following account is intended to serve only as a brief review of the way in which the problem is set up.

The presence of the magnetic field of the earth causes a wave incident on the ionosphere to split into ordinary and extraordinary waves, each of which thereafter travels independently in the ionosphere and is reflected at different heights. To calculate the intensities of the components parallel and perpendicular to the plane of incidence in a downcoming wave, it is necessary to calculate the intensities of these components for the ordinary and extraordinary waves and then to take the vector sum of these two waves to give the total field components.

The following symbols will be used together with Heaviside-Lorentz units; bold face symbols are vectors. A dot over a symbol denotes time differentiation.

e = charge on an electron.
m = mass of an electron.
N = electron density.
$\mathbf{H}^o$ = earth's magnetic field.
v = mean frequency of collisions between free electrons and neutral air molecules.
c = velocity of light.
f = frequency of radio wave.
$\omega = 2\pi f$.
$\mathbf{E}$ = electric field intensity of the radio wave.
$\mathbf{H}$ = magnetic field intensity of the radio wave.
$\mathbf{J}$ = total current (displacement plus convection current).
$\mathbf{V}$ = velocity of electron motion in the ionosphere.
$\mathbf{i}, \mathbf{j}, \mathbf{k}$ = right-handed set of mutually perpendicular unit vectors.

The problem of propagation of radio waves through the ionosphere is of course solved by looking for the appropriate solution of Maxwell's equations.

$$\nabla \times \mathbf{E} = -\frac{1}{c}\dot{\mathbf{H}} \qquad (1)$$

$$\nabla \times \mathbf{H} = \frac{1}{c}\mathbf{J} \qquad (2)$$

In the case of the ionosphere, the free electrons present give rise to a convection current $Ne\mathbf{V}$ so that the total current $\mathbf{J}$ is given by

$$\mathbf{J} = \dot{\mathbf{E}} + Ne\mathbf{V}. \qquad (3)$$

Furthermore the motion of the electrons must satisfy the force equation

$$e\mathbf{E} - m\dot{\mathbf{V}} - vm\mathbf{V} + \frac{e}{c}\mathbf{V} \times \mathbf{H}^o = 0. \qquad (4)$$

Here the term $(e/c)\ \mathbf{V} \times \mathbf{H}^o$ is the force due to the magnetic field of the earth. This term causes the paths of the electrons which are oscillating under the influence of the electric field of the radio wave to be bent (if moving with uniform speed they would be bent into circular helices) and thus causes the wave to be split into ordinary and extraordinary components. The term $vm\mathbf{V}$ is a dissipative force produced by collisions with the neutral air molecules. This term gives rise to the absorption of the wave.

Since we are looking for a plane-wave solution, we substitute into equations (1) to (4) the plane-wave function, $\exp[i(k\mu\ \mathbf{n}\cdot\mathbf{r}-\omega t)]$. Here μ denotes the complex index of refraction, $\mathbf{n}$ denotes a unit vector normal to the wave front, and $\mathbf{r}$ is a vector denoting the position of the field point from a fixed origin. The field equations can be satisfied providing the parameters fulfill certain relations which are

found by substituting the plane-wave function into the equations. Thus μ comes out to be double-valued showing the splitting into ordinary and extraordinary wave components. For each of these values of the index of refraction, the field equations can be solved for the electric and magnetic intensities. When such a solution is carried out for the case of plane waves incident on the ionosphere from a transmitter on the ground, it is found that the ordinary and extraordinary waves are both elliptically polarized, the state of polarization being independent of the polarization of the incident wave. This result is important for short-wave direction finding since it means that no matter what the transmitted polarization may be, the downcoming ionospheric waves will, in general, be elliptically polarized and can therefore be resolved into two plane-wave components polarized parallel and perpendicular respectively to the plane of incidence and having a suitable phase diffecence. The difference in azimuthal response of a direction finder to the several components of such a field will therefore result in inaccuracies of the bearing.

2.2.3 Effects of Fading

These inaccuracies will vary as the state of polarization of the downcoming wave varies, particularly since both the phase and amplitude of the parallel and perpendicular components varies. Norton[3] has treated this problem quantitatively by considering the effect of fading of the ordinary and extraordinary waves on the resultant downcoming wave (equal to their vector sum).

There are two causes for the fading of ionospheric waves: phase interference between waves traveling along slightly different paths in the ionosphere, and changes in the absorption of radio waves caused by variations in the ionization distribution in the ionosphere. The phase interference is responsible for the rapid changes in intensity which occur from minute to minute, while the changes of absorption are responsible for slower changes in the average level of the received fields which occur from hour to hour and from day to day. These latter changes can be neglected for this work.

The fading caused by phase interference is a result of the irregular nature of the ionosphere. The ionosphere probably consists of clouds of ions distributed in such a manner that a single downcoming ionospheric wave actually consists of a large number of component waves, each of which has been reflected at a slightly different place in the ionosphere. These separate wave components, since they have traveled along slightly different paths through the ionosphere, will arrive at the ground with random relative phases. This fading has been found experimentally[4] to follow a distribution law first derived by Rayleigh which gives the resultant of a large number of waves of the same frequency but of arbitrary phase. Using this distribution law, the fading of the ordinary and extraordinary downcoming waves can individually be determined and therefore also the fading of the resultant downcoming wave. Clearly this fading of the ordinary and extraordinary waves gives rise to variations in the state of polarization of the resultant downcoming wave with concomitant variations in the d-f polarization error. Such variations in state of polarization have been observed experimentally and account for the swinging of bearings observed in practice. The complete analysis of the fading problem leads to the general conclusion that, except for the two special cases given below, the relative amplitude and relative phase of the parallel and perpendicular components of a downcoming ionospheric wave will have a random distribution, the distribution being more nearly random the higher the frequency. The average of a series of swinging bearings will then be close to the true bearing (excluding lateral deviation) except for frequencies near the magneto-ionic frequency or near the maximum usable frequency.

Several conclusions relative to direction finding may be drawn from the preceding discussion. It is clear that the state of polarization of downcoming ionospheric waves will vary over wide limits, there being times when the parallel component only is present and times when the perpendicular component only is present. The phase between these two components also can have any value. These variations have been observed experimentally by direct

measurements of the state of polarization of downcoming waves. Busignies[7] has proposed that some of the large bearing errors previously ascribed to lateral deviation may perhaps be accounted for as polarization errors occurring when the desired component of the wave is practically zero. Clearly, for these cases even a direction finder having a very low standard wave error would exhibit a large bearing error. It is therefore evident that the reduction of the standard wave error alone will not prevent the occurrence of a certain percentage of large bearing errors. Since the period when this happens will usually be short, NBS and Busignies have independently proposed a direction finder in which the taking of bearings is automatically prevented unless the state of polarization of the incident radio wave is favorable for accurate bearings. The antenna system in this method is used not only to take bearings but also to measure the relative polarization of the downcoming wave and thus to control the indications of the direction finder.

A direction finder using spaced, horizontal loop-antenna elements has been suggested by NBS[4] and others as having favorable properties for accurate direction finding. The operation of such a direction finder requires a perpendicular component in the downcoming wave. The preceding analysis has shown that such components will be present approximately equally with the parallel components, so that direction finders designed for either component are feasible.

2.2.6 Effect of Ground on Total Field at the Direction Finder

The preceding discussion of the nature of the downcoming waves must be supplemented by a consideration of the effect of the ground reflection on the resultant field at the direction finder. The response of the antenna system in this resultant field can then be calculated and therefore also the polarization errors.

The wave coming down from the ionosphere will be effectively a plane wave since it will have come from a great distance. It will be reflected from the ground, assumed here to be flat and homogeneous, so that the total field will be given by the vector sum of the direct and ground-reflected waves. The reflected wave can be calculated by using Fresnel's equations; some typical cases will be given here to illustrate the large magnitude of the effect on the resultant field. This will have an important bearing on the selection of a suitable figure of merit for polarization error in direction finders. Figure 1 shows the geometry of the problem for the case in which the electric vector lies in the plane of incidence. The electric vector $E_{p,d}$ of the incident downcoming wave is shown as a dotted line while the electric vector $E_{p,r}$ of the corresponding ground-reflected wave is shown as a solid line. Since the vertical component $E_{p,r}$ of the resultant field is in general out of phase with the component parallel to the ground, $E_{p,r}$, the resultant vector E_p will rotate in an ellipse in the i-k plane (the plane of incidence).

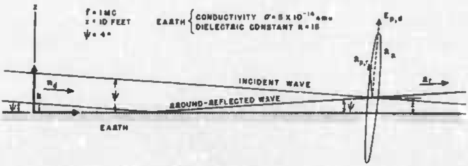


FIGURE 1. Polarization ellipse for electric vector E_p at height z above ground.

Figure 1 has been drawn for the particular case of a downcoming wave on a frequency of 1 mc arriving at an angle of elevation $\psi = 4°$ over land of average conductivity $\sigma = 5 \times 10^{-14}$ emu and with a dielectric constant $K = 15$; the ellipse represents the resultant field at a height $z = 10$ feet.

The general equations for the reflected wave are given by Fresnel's formulas as follows:

$$\mathbf{E}_{p,r} = R_p \mathbf{E}_{p,i} \tag{5}$$

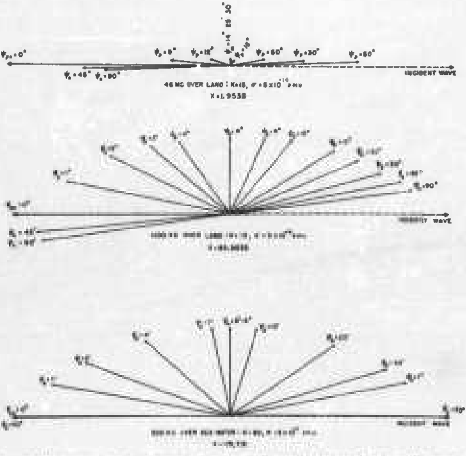
$$\mathbf{E}_{n,r} = R_n \mathbf{E}_{n,i} \tag{6}$$

where the plane-wave reflection coefficients are given by

$$R_p = \frac{n^2 \sin \psi - \sqrt{n^2 - \cos^2 \psi}}{n^2 \sin \psi + \sqrt{n^2 - \cos^2 \psi}} \tag{7}$$

$$R_n = \frac{\sin \psi - \sqrt{n^2 - \cos^2 \psi}}{\sin \psi + \sqrt{n^2 - \cos^2 \psi}}. \tag{8}$$

In these equations n is the complex index of refraction of the earth and $n^2 = K + iX$ where $X = 1.797 \times 10^{11}\sigma/f$; σ is the conductivity of the ground in emu, f is frequency in megacycles per second, and K is the dielectric constant of



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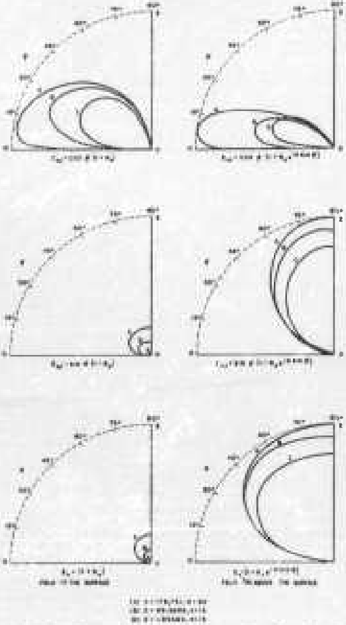

FIGURE 3. Vertical and horizontal components of total field near earth's surface when plane wave of unit intensity is incident at angle θ.

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the ground. The numerical magnitude of the quantity X is of fundamental importance in connection with the effect of the ground on the radio waves, the nature of the reflection being radically different in the two cases when X is very much larger than the dielectric constant and when X is very much smaller than the dielectric constant.

TABLE 1 Values of X for various frequencies and for the ground conductivities normally encountered in practice.

Description	σ(emu)	f(mc)	X
Land, low conductivity	1×10^{-14}	1.6	11.23
Land, average conductivity	5×10^{-14}	1	89.86
Land, average conductivity	5×10^{-14}	5	17.97
Land, average conductivity	5×10^{-14}	46	1.95
Sea water	5×10^{-11}	5	1.79×10^{4}
Sea water	5×10^{-11}	46	1.95×10^{3}
Copper	5.8×10^{-4}	10	1.04×10^{11}

The dielectric constant varies over a much more limited range, from unity for air to 80 for water. The dielectric constant for land varies from about 5 up to about 30. In many of the calculations on specific direction finders given later on in this report, computations were desired for typical cases; accordingly average ground constants of $K = 15$ and $\sigma = 5\times10^{-14}$ emu were assumed.

Figure 2 illustrates the intensity and phase relationship (as given by R_p or R_n) between the incident and the ground-reflected waves for the case $z = 0$. The vector diagrams give the cases in which the electric vector is either parallel or perpendicular to the plane of incidence for three different frequencies and sets of ground constants. The upper diagram corresponds to an ultra-high frequency and ground constants such that $X << K$: the middle diagram corresponds to a frequency in the standard broadcast band and average ground constants; and the bottom diagram corresponds to reflection at 500 kc over sea water. The two upper diagrams therefore show the effect of different frequencies over average land while the two bottom diagrams show the effect of changing the index of refraction at one end of the band. On these diagrams the intensities of the direct and ground-reflected waves are represented by the length of the arrows while the phase of the ground-reflected wave is represented by the angle at which the solid arrow is drawn; the phase of the incident wave is zero in each case. Each of the solid arrows represents the ground-reflected wave for a different angle of incidence of the incident wave.

The resultant wave at a d-f antenna will be the vector sum of the incident and ground-reflected waves. These must be added in proper phase, the phase difference being caused in part by a phase shift introduced on reflection and in part by the difference in path length. When this is done, the total field intensity is found to be given by the following equations:

$$E_{p,z} = E_{p,d} \sin\psi \left| 1 - R_p e^{4\pi i(z/\lambda)\sin\psi} \right| \tag{9}$$

$$E_{p,x} = E_{p,d} \cos\psi \left| 1 + R_p e^{4\pi i(z/\lambda)\sin\psi} \right| \tag{10}$$

$$E_n = E_{n,d} \left| 1 + R_n e^{4\pi i(z/\lambda)\sin\psi} \right|. \tag{11}$$

Figure 3 shows these three components for the case when $E_{p,d} = E_{n,d} = 1$ for heights $z = 0$ and $\lambda/4$ above ground and for the same frequencies and sets of ground constants as for Figure 2. Notice that the vertical component $E_{p,z}$ increases, at $z = 0$, with increasing values of X (increasing conductivity or decreasing frequency) while the horizontal components $E_{p,x}$ and E_n decrease with increasing values of X. On the other hand, at a height $\lambda/4$ above the ground, $E_{p,x}$ and E_n increase with increasing values of X. These figures indicate that a direction finder designed to respond to the E_p component of the wave should be located over ground with the highest possible conductivity to increase $E_{p,z}$ and decrease E_n which is usually responsible for polarization errors.

The relative values of the total field components will depend on the height z at which the fields are considered. Since $E_{p,z}$ is usually the desired field component while E_n is the undesired component, the ratio $|E_n/E_{p,z}|$ is shown in Figures 4 and 5 for frequencies of 2 and 20 mc as a function of height z of the receiving point. These curves are drawn for equal parallel and normal components in the downcoming wave and are given for several ground constants and angles of elevation. The important result to be noted is that, under most conditions, the effect of the ground is to suppress the horizontal component in the resultant wave in comparison to the vertical component. When E_n is the undesired field component, it should be clear from these figures that the d-f antenna

system should be located as near as possible to ground and over ground with the highest possible conductivity. At 20 mc, Figure 5 shows that for large angles of elevation the vertical component rather than the horizontal component is often suppressed. From these two figures it is clear that in the case of a direction

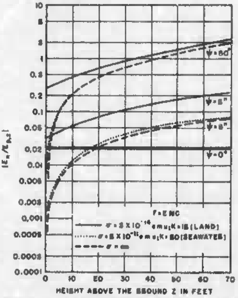

FIGURE 4. Ratio of resultant horizontal to vertical electric field components when plane wave with equal parallel and normal components is incident on ground at angle of elevation ψ.

finder designed to respond to E_n, such as the spaced horizontal loop-antenna type, the antenna system should not be placed too close to the ground. In fact an optimum height of about $\lambda/3$ to $\lambda/4$ is indicated on Figure 5.

Figure 6 indicates the maximum height for direction finders designed to reject E_n. This figure shows the height above perfect ground ($\sigma = \infty$) at which $|E_n/E_{p,t}|$ equals $|E_{n,d}/E_{p,d}|$. Below this height the ground acts to suppress E_n so that the direction finder should be kept below the limit. Over imperfect ground this limiting height will be even less.

So far expressions for the electric field components only have been considered. In the case of loop-antenna direction finders the magnetic field components are also needed and so are given below for the resultant field of an incident and ground-reflected wave using Heaviside-Lorentz units as before.

$$H_{p,x} = E_{n,d} \sin\psi \, |1 - R_n e^{4\pi i (z/\lambda) \sin\psi}| \tag{12}$$

$$H_{p,d} = E_{n,d} \cos\psi \, |1 + R_n e^{4\pi i (z/\lambda) \sin\psi}| \tag{13}$$

$$H_n = E_{p,d} \, |1 + R_p e^{4\pi i (z/\lambda) \sin\psi}| \, . \tag{14}$$

A consideration of these equations indicates the effect of the height above ground when the direction finder uses loop-antenna elements, that

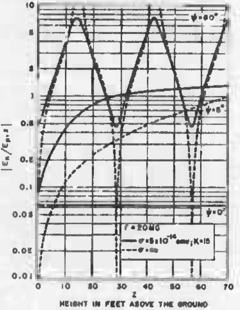

FIGURE 5. Data similar to that in Figure 4 except at frequency of 20 mc.

is, magnetic dipole elements. In particular, for the spaced, vertical, coaxial loop-antenna system, the undesired field component is E_n, while the desired component is H_n. Since

$$E_n/H_n = |E_n/E_{p,d}| \cos\psi \, . \tag{15}$$

Figures 4 and 5 may be used in connection with the spaced coaxial loop-antenna system simply by multiplying the values given in the curves by the factor $\cos\psi$.

2.2.7 The Calculation of Polarization Errors

The preceding sections have shown how the total field components at a direction finder are determined for downcoming ionospheric waves for any frequency or values of ground constants. The response of the antenna system in such a field must next be calculated, including the effect of both the desired and undesired

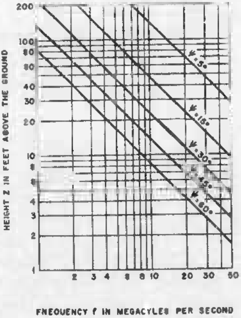

FIGURE 6. Maximum height above perfect ground at which $|E_n/E_{p,z}| = |E_{n,d}/E_{p,d}|$.

field components on the d-f azimuthal directional pattern. The departure of this directional pattern from the ideal pattern obtained for the desired field component alone is the cause of the polarization error of the antenna system. The difference in bearing given by the ideal pattern and the actual pattern is equal to the polarization error. When the incident wave is assumed to have equal parallel and perpendicular components of such a phase relation as to cause maximum bearing error and to have an angle of elevation of 45°, the calculated error will be Barfield's standard wave error.

The response of the antenna system to any field component will be proportional to that component and will have a certain functional dependence on the azimuthal angle ϕ and angle of elevation ψ of the incident wave. The azimuth angle ϕ is the angle between the plane of propagation and the vertical plane passing through the centers of the two spaced antenna elements. The output voltage V of the antenna system can therefore be written as follows, where the voltages induced in the antenna elements and in the feeders are arbitrarily separated.

$$V = V_{\text{antenna}} + V_{\text{feeders}} \tag{16}$$

$$V_{\text{ant}} = h_x E_{p,x} F_x(\phi,\psi) + h_z E_{p,z} F_z(\phi,\psi) + h_y E_n F_y(\phi,\psi) \tag{17}$$

$$V_{\text{feed}} = k_x E_{p,x} f_x(\phi,\psi) + k_z E_{p,z} f_z(\phi,\psi) + k_y E_n f_y(\phi,\psi)\,. \tag{18}$$

In these equations the proportionality constants h and k, corresponding to desired and undesired pickup respectively, are to be determined experimentally. The feeder voltage here is meant to include all undesired voltages. The National Bureau of Standards has adopted a standard value of ψ as zero in this work so that the values of h and k may be determined by measurements on the ground, that is, at horizontal incidence; this seems to be possible for most direction finders. The functions $F(\phi,\psi)$ and $f(\phi,\psi)$ give the directional dependence of each term in the response and are complex quantities including the phase of each term. The functions are dimensionless, while V is to be measured in volts and the field intensities in volts per meter. In this case the constants h will be measured in meters. These functions will depend on the particular antenna system being considered and are used in the preceding equations as holding for a single pair of spaced antenna elements, that is, for a rotatable type of direction finder. Fixed direction finders will be considered later.

Equations (16) to (18) include the effect of the ground reflection. In most cases the functions $F(\phi,\psi)$ and $f(\phi,\psi)$ can be accurately written down a priori from a knowledge of the antenna structure. *The NBS procedure usually used is to assume a priori the dependence on ψ while the dependence on ϕ can be determined by measurements on the ground.* The total field

components $H_{p,x}$, $H_{p,z}$, and H_s could be used in equations (17) and (18) instead of the electric field components with equal generality. The fields can induce voltages in the antenna system directly or indirectly through pickup and reradiation of wires, supporting posts, etc. In any case, the total output voltage can be found as a function of ϕ; that is, the azimuthal directivity pattern will be given by equation (16) and can be rewritten as follows:

$$V = E_{p,z}\left\{(h_xF_x + k_xf_x)\frac{E_{p,x}}{E_{p,z}} + (h_zF_z + k_zf_z) + (h_yF_y + k_yf_y)\frac{E_s}{E_{p,z}}\right\}. \quad (19)$$

Here the shape of the directional pattern is seen to be determined simply by the ratio of the total field components, which in turn are fixed by the ratio of the parallel to perpendicular components in the downcoming wave. Therefore the units used for E or H can be arbitrary when calculating polarization errors since only the ratios of the field components are needed. By setting $E_{s,d} = E_{p,d}$ and $\psi = 45°$, the directional pattern for standard waves can be found. In most direction finders many of the h and k constants appearing in equation (19) are zero or negligibly small; also the different h or k constants are sometimes equal. This simplifies the problem considerably.

Equation (19) gives the phase and amplitude of the output voltage as a function of ϕ. This directional pattern therefore can be compared with the ideal, desired, directional pattern for either the phase-comparison or amplitude-comparison d-f type. The h constants corresponding to the wanted response should be large compared to the k constants corresponding to the unwanted response in order that the distortion of the directional pattern and therefore the polarization error be a minimum. The ratios of h to k can therefore be used as figures of merit for judging the freedom from polarization error of a given direction finder. The use of these ratios for this purpose has the advantage that the ratios are independent of the ground constants and height of the direction finder above the ground. Each type of direction finder will in general require a different procedure to be used in determining the polarization error from equation (19). For example, in those direction finders which determine a bearing by rotating the antenna system until a minimum in the output voltage is found, the bearing, ϕ, will be given by the equation

$$\frac{d}{d\phi}|V| = 0. \quad (20)$$

Since in most rotatable systems the true bearing is given by $\phi = 90°$, the bearing error or polarization error ϵ will then be given by $90° - \phi$ where ϕ is the azimuth angle satisfying equation (20). The value of ϵ obtained from equation (20), or other determining equations depending on the direction finder, will be a function of the phase angle of the various terms in V and these in turn will be variable since the phase of the components in the downcoming wave are random. The maximum value of ϵ is usually the value desired. This is then determined by letting the relative phase of $E_{p,d}$ and $E_{s,d}$ be varied until the maximum value of ϵ is found.

For the case of fixed-type direction finders, the directional pattern is found for each pair of antennas and the bearing determined from these patterns in a manner depending upon the particular d-f type. A common example is the type using a goniometer or similar principle with the antenna pairs at right angles to each other, such as the Western Electric-Civil Aeronautics Administration and International Telephone and Radio fixed direction finders. For this type the *observed* bearing θ relative to the plane through one of the fixed pairs of antennas will be given by

$$\tan\theta = \frac{|V_1|}{|V_2|} \quad (21)$$

where V_1 and V_2 are the output voltages of the two pairs of antennas. The correct bearing ϕ relative to the same plane is given by

$$\tan\phi = \frac{|V_1|}{|V_2|} = \tan\theta \quad (22)$$

only when V_1 and V_2 follow the ideal directional patterns for which the antenna systems were theoretically designed. The polarization error, ϵ, is in this case given by $\phi - \theta$. The maximum value must again be determined by varying the phase of the field components in the downcoming wave.

TYPICAL CALCULATION

To illustrate the method of calculating polarization errors the case of a rotatable, balanced H antenna will be worked out.[6]

The rotatable balanced H-antenna direction finder is a spaced, electric dipole system of the Adcock variety in which the dipoles are differentially connected by means of horizontal transmission lines. These lines are sometimes enclosed by a metal shield and sometimes not. For vertical dipoles, the antenna elements will respond directly only to the vertical electric component of the field of the radio wave. Voltage may be induced in the dipoles by the horizontal component of the field, indirectly, if the coupling to the dipoles of some other part of the system excited by the horizontal field components is not negligible. The polar response pattern of a vertical dipole is nondirectional in azimuth, while it is a figure eight having circular lobes in the vertical plane when the length of the dipole is small compared to the wavelength λ. In general the vertical directional pattern of a dipole depends on its length and height above ground. In the case of the direction finders measured by NBS, the antenna was always short enough to be considered as a pure doublet with a figure eight response pattern, at least for elevation angles ψ up to 45 to 60 degrees. The response of the antenna elements will therefore be taken to be proportional to $E_{p,z}$ alone and the pattern will thus be a figure eight in the vertical plane. The function, $F_z(\phi,\psi)$, however, includes not only the directivity function for a single dipole but also for the two dipoles differentially connected together. The total dipole response will be the vector difference between the voltages induced in the individual dipoles and so will be proportional to twice the sine of half the phase difference between these voltages. This phase difference will be $2\pi(d/\lambda)\cos\psi\cos\phi$ where d is the spacing between the dipoles. Accordingly the total output voltage is proportional to $2E_{p,z}\sin[\pi(d/\lambda)\cos\psi\cos\phi]$ or, when d/λ is small as is generally the case, to $2\pi E_{p,z}(d/\lambda)\cos\psi\cos\phi$. It therefore follows for this case that

$$|F_z(\phi,\psi)| = \cos\psi\cos\phi,$$

or if $\epsilon = 90° - \phi$,

$$|F_z(\epsilon,\psi)| = \cos\psi\sin\epsilon. \qquad (23)$$

The balanced H antenna also has an undesired response to the horizontal components of the field E_n and $E_{p,x}$. The mechanism of this response is not completely understood, except that it is caused by the voltage induced in the horizontal feeders or the shield surrounding the feeders. Clearly the proportionality constant for this pickup will be the same for response to both E_n and $E_{p,x}$. Also the response will be nondirectional in the vertical plane since the feeders can be considered to act as a horizontal antenna (loaded by the dipoles, unless separated, for example, by cathode followers). The following explanations have been proposed for this unwanted response: (1) The system is unbalanced because the lower halves of the dipoles are closer to the ground than the upper halves. (2) The feeders or feeder shield have unbalanced coupling to the dipoles. In either case the directional pattern for this response would be expected to be the same as that of a horizontal doublet (since the length of the feeders is short compared to λ). Finally, the azimuthal response pattern can be determined by measurements and has been found to be that of a horizontal doublet. It follows therefore that

$$|f_y(\epsilon,\psi)| = \cos\epsilon \qquad (24)$$

and

$$|F_x(\epsilon,\psi)| = \sin\epsilon. \qquad (25)$$

Since for this antenna system $h_x = h_y = k_z = 0$ and $k_x = k_y$, the total output voltage, dropping the subscripts, is

$$V = hE_{p,z}\cos\psi\sin\epsilon + kE_{p,x}e^{i\Delta}\sin\epsilon + kE_n e^{i\beta}\cos\epsilon. \qquad (26)$$

Here β is the phase angle between the output voltage induced by E_n and that induced by $E_{p,z}$. It can have any value in practice as already explained. The phase angle Δ is in part made up of a phase shift between $E_{p,z}$ and $E_{p,x}$ introduced by the ground reflection and in part caused by phase shift in the antenna circuit depending upon the differential antenna connection and the exact mechanism of horizontal response.

Equation (26) gives the d-f azimuthal directional pattern. The ideal, desired pattern is the pattern obtained when either k or E_n is zero. This will be a figure eight on a polar diagram with a null at $\epsilon = 0$, the true bearing. In equa-

tion (26) there will not be a null but simply a minimum response unless $\beta = 0$. For this case ($\beta = 0$) the figure eight will be rotated so that the null does not occur for $\epsilon = 0$. Since the bearing is taken as that value of ϵ for which $|V|$ is a minimum, an incorrect bearing will be obtained. In general, the response pattern will not have a null but a broad minimum and in addition will be rotated. By solving the equation $\partial|V|/\partial\epsilon = 0$, the bearing error or polarization error can be obtained (ϵ equals the bearing error since the true bearing is given by $\epsilon = 0$). It is found that the error is a maximum when $\beta = 0$, that is, for cophased *output* voltages. The *maximum* polarization error is given by

$$\tan \epsilon = \frac{|kE_s|}{|hE_{p,t}\cos\psi + kE_{p,z}e^{i\Delta}|}. \qquad (27)$$

For downcoming ionospheric waves, the values of $E_{p,s}$, $E_{p,z}$, and E_s can be obtained from equations (9), (10), and (11). It is clear that only the ratios of the field components need be known. Also only the pickup ratio h/k need be known as seen by rewriting equation (27) as

$$\tan \epsilon = \frac{1}{\left|\frac{hE_{p,t}}{kE_s}\cos\psi + \frac{E_{p,z}}{E_s}e^{i\Delta}\right|}. \qquad (28)$$

In general, the h and k constants only need be measured to within a constant factor since this constant can always be factored out of the expression for V and so will not affect the shape of the directivity pattern.

By taking $E_{p,t} = E_{s,t}$ and $\psi = 45°$, equation (27) gives the value of the standard wave error of the balanced H antenna. However, the polarization error for all values of ψ can be obtained once h/k is known.

In general, the polarization error is smaller, the larger the pickup ratio h/k, as seen from equation (28). The National Bureau of Standards has proposed the use of the pickup ratio as a figure of merit for polarization error in direction finders. It is clear that the pickup ratio is independent of the ground constants and of the height of the direction finder above the ground and so lends itself to the comparison of different direction finders of the same type, that is, following the same law for polarization error, such as equation (28). This comparison as to accuracy can therefore be separated from the complicating influence of the ground and height above ground. Once this fundamental constant is known, not only the standard wave error but the polarization errors for all values of ψ can be determined for any particular ground and antenna height. A curve of ϵ vs ψ can be plotted; it is this complete curve which should be compared with similar curves of other *types* of direction finders to compare their accuracy relative to polarization errors. The pickup ratio also furnishes a figure of merit by which the progress of development work on a particular direction finder can be judged. The effect of changes in the design can thus be studied and the cause and mechanism of polarization errors isolated.

A useful figure of merit somewhat similar to the standard wave error is the polarization error for a *horizontally* incident wave with equal $E_{p,t}$ and E_s components of each phase as to cause maximum bearing error. The error for this wave will be called the horizontal wave error, ϵ_0, while Barfield's standard wave error will be denoted by ϵ_{45}. For the cases of the rotatable H-antenna direction finder, $\tan \epsilon_0 = k/h$. The ϵ_0 error is independent of the ground constants or height of the direction finder above the ground.

2.3 MEASUREMENT OF POLARIZATION ERROR

2.3.1 Plane Wave Measurements

The preceding section has shown that the problem of the measurement of polarization error can be reduced to that of measuring the pickup factors, the constants h and k. In most direction finders the response of the antenna system can be reduced to a single term by placing the antenna in a suitable plane-wave radiation field, having only one component such as E_s or $E_{p,s}$, and orienting the antenna to the proper azimuth angle. By properly choosing the field, the antenna output voltage can then be made to be

$$V_1 = hE_1F(\phi,\psi) \qquad (29)$$

or

$$V_2 = kE_2f(\phi,\psi) \qquad (30)$$

where E_1 and E_2 are the particular field components used. The field intensity can be meas-

ured by means of a field-intensity meter, and the output voltage V of the antenna system also determined. The value of $F(\phi,\psi)$ or $f(\phi,\psi)$ is known, so that equations (29) and (30) can be solved for the pickup factors.

$$h = \frac{V_1}{E_1 F(\phi,\psi)} \qquad (31)$$

$$k = \frac{V_2}{E_2 f(\phi,\psi)} \qquad (32)$$

Usually the field used is one at horizontal incidence so that $\psi = 0$. This simplifies the technique by allowing all measurements to be made close to the ground. Often the measurements *must* be made at horizontal incidence to reduce the response of the antenna to a single term; the presence of voltage corresponding to more than one term would require a knowledge of the phase of each term. Also the measurements are usually made at particular values of ϕ in order to obtain various experimental advantages. The pickup factors can be defined as the output voltage per unit field intensity for azimuth and elevation angles such that $F(\phi,\psi)$ and $f(\phi,\psi)$ are unity (provided they can assume such values, as is usually the case). This is the reason for the designation pickup factor.

Usually the azimuthal directional or response patterns of the system are determined by measuring V and E for the special fields already indicated as a function of ϕ (with $\psi = 0$). If the response is defined as the ratio of V to E, that is, the output voltage per unit field intensity, these curves will be given by

$$\frac{V_1}{E_1} = hF(\phi,0)$$

or

$$\frac{V_2}{E_2} = kf(\phi,0).$$

The response is equal to the pickup factors if and when F or f is equal to unity.

To illustrate the procedure just outlined, the case of the balanced H antenna will again be treated. When the antenna is placed in the field of a plane wave polarized parallel to the plane of incidence so that $E_{p,y} = E_n = 0$, the output voltage will be

$$V = E_{p,z} \cos \psi \sin \epsilon. \qquad (33)$$

When $\psi = 0$ and $\epsilon = 90°$, V and $E_{p,z}$ are measured, giving $h = V/E_{p,z}$ in meters if V is measured in volts and $E_{p,z}$ in volts per meter. Similarly if a plane wave polarized perpendicular to the plane of incidence is used so that $E_{p,x} = E_{p,z} = 0$, the output voltage will be

$$V = kE_n \cos \epsilon, \qquad (34)$$

so that $k = V/E_n$ when $\epsilon = 0$. In this case the pickup factors are equal to the output voltages per unit field intensity at maximum response.

2.3.2 Application to Buried U Direction Finders

A difficulty arises when applying the method just outlined to the buried U-antenna direction finder. As its name indicates, the antenna consists of vertical electric elements connected by horizontal feeders or transmission lines which are buried below the surface of the earth. By this means the field intensity at the feeders is greatly reduced both because of the partial reflection of the incident wave by the ground and the attenuation of the transmitted wave by absorption in the ground. The expression for the output voltage of this antenna system in the field of a downcoming ionospheric wave will consist of terms for the voltage induced in the antenna elements and terms for the voltage induced in the feeders. The voltage induced in the antenna elements will involve the desired pickup factor h and the field intensity at the antenna elements, while that induced in the feeders will involve the undesired pickup factor k and the field intensity at the feeders. Since the pickup factor k is the proportionality constant relating the output voltage of the antenna as a result of voltage induced in the feeder to the field intensity at the feeders, both the output voltage and field intensity must be measured to determine k. Because the feeders are buried, the difficulty then arises of measuring the field intensity below the ground. This difficulty can be met, however, by a procedure to be described in which expressions are used for the field intensity at any depth Δ below the surface of ground having arbitrary constants.

The procedure for avoiding the above difficulty in measuring k is as follows. In all methods of studying polarization errors, the errors must be determined from a knowledge of the field intensities in an *incident*, downcoming wave. In the NBS method the output voltage of the antenna system must be calculated when these field intensities are given. This involves

the calculation of the field intensity at the antenna components when the field intensity in the *incident* wave is given. However, the field intensity at the antenna components can be specified in terms of the field intensity at any other point. If this reference point is taken to be the same for the vertical antenna elements and for the buried feeders and to be above ground, both h and k can be measured by procedures quite similar to those already given. The reference point could be taken at the center of the d-f antenna system at a height, z, above the ground.

Before proceeding to a development of this procedure some specific points must be considered concerning the expression for the output voltage of the vertical antenna elements and concerning the effect of a ground mat. In the other direction finders which have been considered, namely those using elevated antennas, the field intensity was taken to be the same over the entire region occupied by the antenna elements. This assumption is probably not a good enough approximation for the buried U antenna, so that an integration over the antenna elements would be required to obtain a more accurate expression for the induced voltage. However, in the simplified analysis to be given, this will not be done, the assumption being made that the field intensity at the center of the antenna system can be used for computing the voltage induced in the antenna elements. Some buried U-antenna systems use a large ground mat which must be considered when computing the field intensities above and below the surface of the ground. However, in the analysis to be given it will be assumed that no ground mat is present or that it is so small compared to the wavelength that its effect on the field intensity can be neglected.

A derivation will now be given of the expression for the total output voltage of the buried U antenna without ground mat as a function of incident angle as the principal parameter. The polarization errors can then be derived according to the particular indicating method used and so will not be given here. The voltage induced in the feeders by $E_{p,t}$ can be safely neglected since $E_{p,t}$ will be very small except, perhaps, for very large angles of elevation of the incident wave. The output voltage as a result of voltage induced in the antenna elements will be $hE_{p,t} \cos\psi \sin\epsilon$, where $E_{p,t}$ is the vertical electric field intensity at the center of the antenna system (a height z above the ground). The output voltage as a result of voltage induced in the feeders will be $kE_{n,t} \cos\epsilon$, where $E_{n,t}$ is the transmitted horizontal electric field intensity at the depth, Δ, below the ground where the feeders are buried. From the material below on d-f sites the value of $E_{n,t}$ is taken as

$$E_{n,t} = E_{n,d}\left|(1+R_n)\, e^{(2\pi/\lambda)(z\sin\psi + \Delta\sqrt{n^2-\cos^2\psi})}\right| \qquad (35)$$

while $E_{p,t}$ is given by equation (10) as

$$E_{p,t} = E_{p,d}\cos\psi\left|1 + R_p e^{4\pi i (z/\lambda)\sin\psi}\right|. \qquad (36)$$

However, the field intensity E_n at the height z above the ground was given by equation (11) as

$$E_n = E_{n,d}\left|1 + R_n e^{4\pi i (z/\lambda)\sin\psi}\right|. \qquad (37)$$

Therefore, if the direction finder is placed in the field of a perpendicularly polarized downcoming wave in order to measure k, E_n can be measured at the height z and $E_{n,d}$ calculated by means of equation (37). Using this value of $E_{n,d}$ the value of $E_{n,t}$ at the depth Δ can be determined by using equation (35). The value of k can then be found from the equation

$$k = \frac{V_n}{E_{n,t}\cos\epsilon} \qquad (38)$$

where V_n is the measured output voltage. It is clear that this procedure effectively measures k in terms of the field intensity at the feeders by measuring the field intensity above the ground and then calculating the field intensity at the depth Δ from this measurement. To do this, the ground constants must be known. However the constant k still is independent of the ground constants or the depth of the feeders below the ground and so is a useful figure of merit for measuring undesired pickup. Once k is measured, the output voltage of the antenna system for any downcoming ionospheric wave will be given by

$$V = hE_{p,t}\cos\psi\sin\epsilon + kE_{n,t}\cos\epsilon\, e^{i\beta} \qquad (39)$$

with $E_{p,t}$ and $E_{n,t}$ given by equations (36) and (35) respectively. Here β is the arbitrary phase angle already discussed in connection with equation (26).

When measuring k in practice, a local transmitter is used which does not generate plane

waves. In this case the fields E_n and $E_{n,t}$ are given below in this report and in equation (194), page 64, of the Norton report.[a] The pickup factor k is then determined by the procedure outlined except that these equations for a local transmitter must be used rather than equations (35) and (37). After finding k by using the equations for a *local transmitter*, equations (35) and (39) must still be used for calculating the polarization error for *downcoming ionospheric waves*. Equations (35) and (37) can be used to determine k even when a local transmitter is used provided that the transmitter is at a great enough distance from the direction finder and at an angle of elevation large enough so that equations (35) and (37) are valid. This point is discussed in detail below. When the local transmitter is used near the ground, however, the exact expressions for the field from a local transmitter must be used to determine k by measurements above the surface of the ground.

3.3.3 Local Transmitter Measurements

The pickup factors h and k which determine the response of a direction finder were defined for plane waves such as downcoming ionospheric waves. The procedures in which special plane waves are used to make possible the measurement of h and k must be modified in practice since the only practicable means of generating such special fields is by the use of a local transmitter placed a relatively short distance from the direction finder. The wave from such a transmitter will simulate an ionospheric wave only approximately, thus introducing into the experimental technique several difficulties which must now be considered.

Two methods of determining polarization errors have been introduced by NBS and RCA[4] respectively.[b] The NBS method uses a local transmitter near the ground while RCA uses an elevated transmitter to generate the fields required in the two methods; accordingly the first method is a horizontal incidence method while the second utilizes downcoming waves. In both methods the two special waves generated are, first, a wave polarized parallel to the plane of incidence so that $E_n = 0$ and, second, a wave polarized perpendicular to the plane of incidence so that $E_p = 0$. In the NBS method the wave of parallel polarization arrives at horizontal incidence so that $E_{p,z} = 0$ also. In practice these conditions are only approximately met, the deviations from the desired fields being as follows.

The $E_{p,z}$ Wave Component

In the NBS method the presence of the ground causes a wave tilt which gives rise to a small $E_{p,z}$ component. The wave tilt is usually less than 10° so that this component can be neglected in practice, especially since it induces voltage which is small in comparison to that of the $E_{p,z}$ component. The pickup factor for the $E_{p,z}$ component is small because it is usually the feeders which are responsible for such pickup.

Generation of Pure Fields

It is very difficult to generate a wave polarized perpendicular to the plane of incidence without also generating some parallel component. The stray parallel component becomes relatively more important, the greater the distance of the local transmitter from the d-f site, since the ground very rapidly attenuates the perpendicular component in comparison to the low attenuation of the parallel component. The more accurate the direction finder, if designed to reject the perpendicular component and to respond to the parallel component, the greater must be the purity of the field to measure the smaller polarization error which such an improved direction finder would have.

The Adcock type of direction finder, in which spaced electric monopole or dipole antenna elements are balanced against each other, relaxes the stringent conditions for purity of the field since the response of the system to E_n can be measured with the antenna oriented at the null position for E_p (and vice versa, although the

[b] The RCA method of measuring polarization errors differs from the NBS method already outlined as follows. The special fields used are those of *downcoming* waves, first polarized parallel to the plane of incidence, then perpendicular to the plane of incidence. The output voltage of the direction finder is measured for each wave and will be V_p and V_n respectively when oriented for maximum response. The polarization error ε in the field of both waves will be given by $\tan \varepsilon = V_n/V_p$.

problem of generating a field with E_n negligible is not difficult). For this reason NBS has measured the pickup factors of such direction finders with the antenna system oriented in the proper null position.

Careful design of the local transmitter helps to prevent the generation of undesired field components. The antenna should contain or be an extension of the shield containing the oscillator and batteries so that current flow will be possible in the desired paths only.

Finally, a flat homogeneous site should be used when making measurements of polarization error.

2.2.4 Field Generated by a Local Radiator

The Surface Wave Component

The presence of the local transmitter near the ground results in the generation of a surface wave component in the wave in addition to the direct and ground-reflected waves.[3] The expressions for the field generated by a local transmitter were also obtained by Burrows,[9-15] who used a somewhat different terminology from that used here. The surface wave terminology will be used in this report since much of the work was carried out by using the equations and methods of K. A. Norton. The vector sum of the direct and ground-reflected waves is called the space wave. The space wave is the only wave present at a direction finder for downcoming ionospheric waves, so that the surface wave component prevents the simulation of such waves by the use of a local transmitter. The presence of the surface wave introduces no difficulty in the NBS method since the total field intensity is measured, the effect of the surface wave thus being allowed for. However, in the RCA method and in Barfield's method, the response of the antenna system will not be the same as for an ionospheric wave arriving at the same angle of elevation. The magnitude of this effect increases as the distance to the transmitter decreases and the angle of elevation decreases; it can be very large for the usual experimental setup. If it is assumed that the surface wave is negligible when it has an intensity less than 1 per cent of the space wave, then, considering the parallel electric field radiated by a vertical electric dipole, it is found that transmissions designed to simulate ionospheric wave transmission must be made from a distance of the order of 2λ when $\psi = 45°$, 60λ when $\psi = 15°$, and 500λ when $\psi = 5°$.

The practical importance of the surface wave component for polarization measurements using elevated transmitters can be illustrated by the experience of the RCA group. In the RCA method, the maximum response of the antenna system to the parallel polarized field and to the perpendicularly polarized field was measured, the ratio being V_p/V_n. Clearly the pickup ratio h/k can be determined from this measurement at the angle ψ, just as the response at the angle ψ can be calculated from the measured values of the pickup ratio. If the response of the antenna system to $E_{p,z}$ is neglected it follows that

$$V_p = hE_{p,x}\cos\psi \tag{40}$$

$$V_n = kE_n \tag{41}$$

since the *maximum* output voltages are measured. The total field components will be proportional to the corresponding field components in the incident wave $E_{p,d}$ and $E_{n,d}$. Thus

$$V_p = hE_{p,d}f_p\cos\psi \tag{42}$$

$$V_n = kE_{n,d}f_n. \tag{43}$$

Here the functions f_p and f_n are given simply by the laws of plane-wave reflection *when ionospheric waves are considered.* Thus

$$f_p = \cos\psi\,[1 + R_p\, e^{4\pi i(z/\lambda)\sin\psi}] \tag{44}$$

$$f_n = 1 + R_n\, e^{4\pi i(z/\lambda)\sin\psi} \tag{45}$$

Putting the proper values of the parameters in these equations and using the measured value of V_p/V_n, the pickup ratio can be solved for, giving

$$\frac{h}{k} = \frac{V_p E_{n,d} f_n}{V_n E_{p,d} f_p}\,\frac{1}{\cos\psi} \tag{46}$$

In this manner RCA determined h/k for values of ψ from near zero up to almost 45°. The pickup ratios thus found were constant for large values of ψ but much greater at low angles than at high. However, when the functions f_p and f_n were computed using the surface wave component as well as the space wave, the pickup ratios thus determined were constant for all values of ψ. Accordingly, the field did not simulate that of an ionospheric wave until elevation

angles of 20° to 30° were reached. If the transmitter were moved further from the direction finder the surface wave would have been reduced since it is attenuated faster than the space wave.

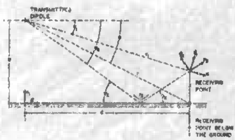

FIGURE 7. Elevated dipole transmitting over finitely conducting ground.

FIELD GENERATED BY VERTICAL AND HORIZONTAL ELECTRIC AND MAGNETIC DIPOLES

For purposes of reference and for use in the next section of this report the complete equations are given for the field from vertical and horizontal electric and magnetic dipoles at distances, d, greater than the wavelength.[5] These expressions refer to dipoles transmitting over a finitely conducting ground as shown in Figures 1 and 7 and include the surface wave component. E_{oe} and H_{oe} are the values of the electric and magnetic radiation fields at a unit distance in free space in the equatorial plane of the electric dipole, while E_{om} and H_{om} are the corresponding values for a magnetic dipole. The $\mathbf{d}\mathbf{k}$ plane is taken as the plane of incidence with $\mathbf{k}$ vertical; Figure 7 defines r_1, r_2, ψ_1, and ψ_2 and the unit vectors

$$\boldsymbol{\psi}_1 = \cos\psi_1 \mathbf{k} + \sin\psi_1 \mathbf{d}, \tag{47}$$

$$\boldsymbol{\psi}_2 = \cos\psi_2 \mathbf{k} - \sin\psi_2 \mathbf{d}, \tag{48}$$

$$\mathbf{r}_1 = \cos\psi_1 \mathbf{d} - \sin\psi_1 \mathbf{k}, \tag{49}$$

$$\mathbf{r}_2 = \cos\psi_2 \mathbf{d} + \sin\psi_2 \mathbf{k}. \tag{50}$$

The expressions for the fields are (here $k = 2\pi/\lambda$):

Vertical Electric and Magnetic Dipoles

$$\mathbf{E} = iE_{oe}\Big\{\cos\psi_1 \frac{e^{ikr_1}}{r_1}\boldsymbol{\psi}_1 + \cos\psi_2 R_p \frac{e^{ikr_2}}{r_2}\boldsymbol{\psi}_2 + \cos\psi_2 (1 - R_p) f(P_e,B_e) \cdot \left[\cos\psi_2 \mathbf{k} + \frac{\sqrt{n^2 - \cos^2\psi_2}}{n^2}\mathbf{d}\right]\frac{e^{i(\phi + kr_2)}}{r_2}\Big\}, \tag{51}$$

$$\mathbf{H}_n = iH_{oe}\boldsymbol{\theta}\Big\{\cos\psi_1 \frac{e^{ikr_1}}{r_1} + \cos\psi_2 R_p \frac{e^{ikr_2}}{r_2} + \cos\psi_2 (1 - R_p) f(P_e,B_e)\frac{e^{i(\phi + kr_2)}}{r_2}\Big\}, \tag{52}$$

$$\mathbf{E}_n = E_{om}\boldsymbol{\theta}\Big\{\cos\psi_1 \frac{e^{ikr_1}}{r_1} + \cos\psi_2 R_n \frac{e^{ikr_2}}{r_2} + \cos\psi_2 (1 - R_n) f(P_m,B_m)\frac{e^{i(\phi + kr_2)}}{r_2}\Big\}, \tag{53}$$

$$\mathbf{H}_p = H_{om}\Big\{\cos\psi_1 \frac{e^{ikr_1}}{r_1}\boldsymbol{\psi}_1 + \cos\psi_2 R_n \frac{e^{ikr_2}}{r_2}\boldsymbol{\psi}_2 + \cos\psi_2 (1 - R_n) f(P_m,B_m) \cdot \left[\cos\psi_2 \mathbf{k} + \sqrt{n^2 - \cos^2\psi_2}\,\mathbf{d}\right]\frac{e^{i(\phi + kr_2)}}{r_2}\Big\}. \tag{54}$$

In this case $\mathbf{E}_n = \mathbf{H}_p = 0$ for the vertical electric dipole while $\mathbf{E}_p = \mathbf{H}_n = 0$ for the vertical magnetic dipole.

Horizontal Electric and Magnetic Dipoles.

Figure 8 illustrates these cases, the electric dipole, and axis of the magnetic dipole (loop) pointing in the positive $\mathbf{i}$ direction.

$$\mathbf{E}_n = iE_{oe}\sin\theta\boldsymbol{\theta}\Big\{\frac{e^{ikr_1}}{r_1} + R_n \frac{e^{ikr_2}}{r_2} + (1 - R_n) f(P_m,B_m)\frac{e^{i(\phi + kr_2)}}{r_2}\Big\}, \tag{55}$$

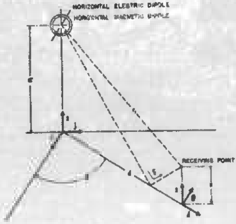

FIGURE 8. Case of horizontal electric and magnetic dipoles.

$$\mathbf{E}_v = iE_{ev}\cos\theta\left\{\sin\psi_1\frac{e^{ikr_1}}{r_1}\boldsymbol{\psi}_1 + \sin\psi_2 R_v\frac{e^{ikr_2}}{r_2}\boldsymbol{\psi}_2\right.$$
$$+\frac{\sqrt{n^2-\cos^2\psi_2}}{2}(1-R_v)f(P_e,B_e)\cdot$$
$$\left.\left[\cos\psi_2\mathbf{k}+\frac{\sqrt{n^2-\cos^2\psi_2}}{n^2}\mathbf{d}\right]\frac{e^{i(\phi+kr_2)}}{r_2}\right\}, \quad (56)$$

$$\mathbf{H}_p = iH_{ev}\sin\theta\left\{\frac{e^{ikr_1}}{r_1}\boldsymbol{\psi}_1 + R_n\frac{e^{ikr_2}}{r_2}\boldsymbol{\psi}_2\right.$$
$$+(1-R_n)f(P_m,B_m)\cdot$$
$$\left.\left[\cos\psi_2\mathbf{k}+\sqrt{n^2-\cos^2\psi_2}\,\mathbf{d}\right]\frac{e^{i(\phi+kr_2)}}{r_2}\right\}, \quad (57)$$

$$\mathbf{H}_n = -iH_{ev}\cos\theta\theta\left\{\sin\psi_1\frac{e^{ikr_1}}{r_1} + \sin\psi_2 R_v\frac{e^{ikr_2}}{r_2}\right.$$
$$\left.+\frac{\sqrt{n^2-\cos^2\psi_2}}{n^2}(1-R_v)f(P_e,B_e)\frac{e^{i(\phi+kr_2)}}{r_2}\right\}, (58)$$

$$\mathbf{E}_p = E_{em}\sin\theta\left\{\frac{e^{ikr_1}}{r_1}\boldsymbol{\psi}_1 + R_n\frac{e^{ikr_2}}{r_2}\boldsymbol{\psi}_2\right.$$
$$+(1-R_n)f(P_e,B_e)\cdot$$
$$\left.\left[\cos\psi_2\mathbf{k}+\frac{\sqrt{n^2-\cos^2\psi_2}}{n^2}\mathbf{k}\right]\frac{e^{i(\phi+kr_2)}}{r_2}\right\}, \quad (59)$$

$$\mathbf{E}_n = E_{em}\cos\theta\theta\left\{\sin\psi_1\frac{e^{ikr_1}}{r_1} + \sin\psi_2 R_n\frac{e^{ikr_2}}{r_2}\right.$$
$$\left.+\sqrt{n^2-\cos^2\psi_2}(1-R_n)f(P_m,R_m)\frac{e^{i(\phi+kr_2)}}{r_2}\right\}, \quad (60)$$

$$\mathbf{H}_n = -H_{em}\sin\theta\theta\left\{\frac{e^{ikr_1}}{r_1}\boldsymbol{\psi}_1 + \sin\psi_2 R_n\frac{e^{ikr_2}}{r_2}\boldsymbol{\psi}_2\right.$$
$$+\sqrt{n^2-\cos^2\psi_2}(1-R_n)f(P_m,B_m)\cdot$$
$$\left.\left[\cos\psi_2\mathbf{k}+\sqrt{n^2-\cos^2\psi_2}\,\mathbf{d}\right]\frac{e^{i(\phi+kr_2)}}{r_2}\right\}. \quad (61)$$

In these equations, R_v and R_n are the plane wave reflection coefficients as already defined. The third term in each equation represents the surface wave and $f(P,B)e^{i\phi}$ is the surface wave attenuation function which is given graphically as a function of P and B in Figures 9 and 10. Here the angle ϕ is the phase of the surface wave attenuation function (not to be confused with the azimuth angle).

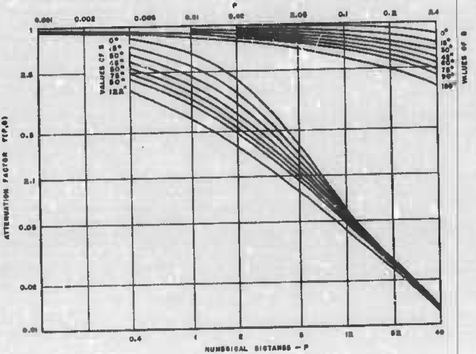

FIGURE 9. Surface wave attenuation function.

$$f(P,B)\, e^{i\phi} = 1 + i\sqrt{\pi P_1}\, e^{-P_1}\operatorname{Erfc}\left(-i\sqrt{P_1}\right), \qquad (62)^{c}$$

$$P_1 = P e^{iB}, \qquad (63)$$

$$P_r\, e^{iB_r} = \frac{ikr_1}{2\cos^2\psi_2}\left[\sin\psi_2 + \frac{\sqrt{n^2 - \cos^2\psi_2}}{n^2}\right]^2, \qquad (64)$$

$$P_m\, e^{iB_m} = \frac{ikr_2}{2\cos^2\psi_2}\left[\sin\psi_2 + \sqrt{n^2 - \cos^2\psi_2}\right]^2. \qquad (65)$$

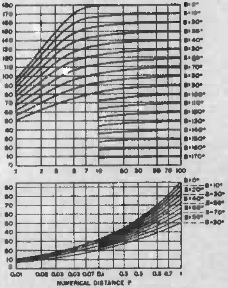

FIGURE 10. Phase of surface wave attenuation function.

P_r and P_m are called numerical distances. The preceding expressions for the field from a local radiator reduce to the values given for the waves when r is allowed to increase without limit.

To indicate the magnitude of the surface wave, Figure 11 shows the ratio of surface to space wave intensities at the surface of the earth radiated from a vertical electric dipole at a height a. It is seen that the surface wave falls off with increasing distance r and increasing elevation angle ψ. For a local transmitter at a distance of one wavelength, very large values of ψ are needed to simulate a downcoming plane wave.

[c] Erfc(x) represents the so-called error function.[11-14]

2.3.8 Radiator Parallax

An examination of equation (56) for the field from a local horizontal electric dipole transmitter reveals that $E_p \neq 0$ except in the equatorial plane where $\cos\phi = 0$. When such a radiator is used for determinations of polarization errors there will be E_p components at the d-f antenna elements since these will lie on either side of the equatorial plane of the transmitting dipole. Furthermore the phase of the fields at the two antenna elements will be opposite so that the induced voltages will add up in the output of the system, as a result of the differential connection of the direction finder, causing an error which will be called "radiator parallax." The response of the antenna system to these E_p components is not desired when using such a horizontal dipole, since the response to E_n alone must be measured to make possible an accurate determination of the corresponding pickup factor.

These undesired parallel components will be present for both horizontal incidence and elevated transmitter methods of measuring polarization error. The presence of the ground is responsible for this state of affairs in the case of the horizontal incidence method where the transmitting and receiving antennas are at the same height above ground, since there is no $E_{p,1}$ component in the direct wave ($\sin\psi_1 = 0$) while there is such a component in the direct wave from an elevated transmitter. The significance of the undesired E_p component in direction-finder testing was first pointed out by W. H. Wirkler of the Collins Radio Company. This component is important because, although small compared to E_n, it is not attenuated so rapidly and so, at large distances from the transmitter, it becomes relatively large enough to render the measurements inaccurate. This holds especially for direction finders having low values of polarization error. *If E_n is generated by means of a vertical magnetic dipole (horizontal loop-antenna), $E_n = 0$ so that this error is avoided.*

When a local transmitter is used to test a spaced vertical loop-antenna direction finder for response to E_n, there will be unwanted H_n components at the two loop antennas if the transmitter uses a horizontal electric dipole.

This case is similar to the one just discussed. This result is seen in equation (58) which shows that the phase of H_x will be opposite at the two loops, thus causing an apparent response to E_x which is misleading. Here also the remedy is to use a horizontal loop antenna in the transmitter.

Situations similar to that just described arise when generating a wave polarized parallel to the plane of incidence. This may also be seen from an examination of the preceding equations (60) and (61). Accordingly the general rule follows that a local transmitter using a vertical electric dipole should be used for generating the wave polarized parallel to the plane of incidence and one using a horizontal loop antenna for generating the wave polarized perpendicular to the plane of incidence.

Collector Parallax

Another error which occurs with local transmitters was also pointed out by the Collins Radio Company. This error, called parallax error, occurs when measurements of polarization error are made on spaced, vertical loop-antenna direction finders.

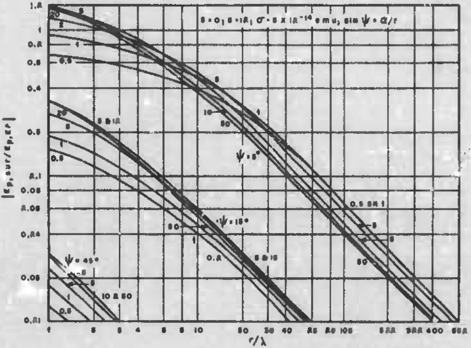

FIGURE 11. Ratio of surface to space wave intensities at surface of earth radiated from vertical electric dipole at height a for frequencies in megacycles for which curves are labeled.

When the response of the antenna system to E_x is tested, the forward tilt of the H_y component of the field results in pickup in the loop antennas which is not balanced out. This effect can be seen by an inspection of equations (57) and (54) which shows an $H_{y,r}$ component of the field, whether a horizontal electric dipole or a horizontal loop antenna is used in transmission. The $H_{y,r}$ component induces voltage in the loop antennas because of the finite "parallax"

angle subtended by the direction finder at the transmitter. This magnetic field component is not parallel to the plane of the H_p loop antennas because the loop antennas lie on either side of the line connecting the center of the direction finder and the center of the transmitter. Furthermore, the pickup in each loop antenna is of opposite phase so that the induced voltages add up.

3.4 EXPERIMENTAL TECHNIQUE

One of the principal results of this work was the development of a method for measuring polarization errors and of experimental techniques to use the method. Representative direction finders were examined by these techniques, it being necessary to find means of overcoming experimental difficulties when the method was applied to particular d-f systems.

The d-f systems examined were: (1) an experimental rotatable, balanced H antenna built and installed at Laurel, Md.—a direction finder using unshielded, horizontal feeders but otherwise practically the same as a Navy DY also installed at Laurel; (2) SCR-551-T1 installed at Fort Monmouth, N. J.; (3) a Western Electric [WE] direction finder developed for the Civil Aeronautics Administration [CAA] and located at LaGuardia Airport, New York City—a ten-frequency, fixed, tuned, Adcock arrangement using balanced H antennas; (4) an elevated, rotatable, spaced loop-antenna system procured from United Air Lines and installed at Laurel, Md.—the system examined with the antennas in both a vertical and a horizontal position; (5) a Collins CXAL direction finder measured at Cedar Rapids, Iowa.

The general procedure will be given here, special procedures being described in the section to which they apply.[14] In all cases except the Collins measurements, a local transmitter employing an electric dipole was used to generate the desired fields. Waves at horizontal incidence were used; first a wave polarized parallel to the plane of incidence, then one polarized perpendicular to the plane of incidence. The error introduced by using a horizontal electric dipole rather than a horizontal loop antenna was not fully appreciated at the time the measurements were made. However, it seems that the conclusions drawn from the measurements would not be materially altered since the polarization errors found were quite large except for the loop-antenna direction finders. The measurements of the Collins installation were carried out by means of a horizontal loop-antenna radiator. Also radiator parallax was negligible in the case of the measurements of the spaced horizontal loop-antenna system.

The results of some preliminary calculations on radiator parallax using the exact equations already given, indicate agreement with the few experimental observations so far available. The errors caused by radiator parallax can be considerable for both high and low elevations of the transmitter. The error decreases with increasing frequency.

The adjustment of the horizontal dipole used for transmission was very critical. Any slight tilt from the horizontal gave rise to a vertical field component which assumed large relative proportions. The rapid attenuation of the horizontal component was responsible for this. It was found that the presence of personnel near the transmitter also caused the generation of undue amounts of unwanted vertical field component. These difficulties were solved by arranging to control the tilt of the dipole by means of cords. For the case of the balanced H antenna, the antenna was oriented for maximum response to E_y and the transmitting dipole rotated until minimum output was indicated. The transmitting dipole was then exactly horizontal; the purity of field at the direction finder was thus determined by the direction finder itself in the equatorial plane of the transmitting dipole.

The output of the antenna system was measured by substituting a standard voltage generator for the antenna and determining the voltage required to give the same output as obtained with the antenna.

The measurement of field intensity was subject to inaccuracies resulting from the presence of the direction finder. However, when the antenna elements were disconnected, the effect on the field intensity was reduced to a negligible value in most cases as shown by the effect of rotating the direction finder. Before disconnecting the antenna elements, the aparent field intensity veried greatly as the direction finder

was rotated, but this variation become negligible after disconnecting the antennas. For the rotatable H antennas it was found that disconnecting the dipoles was not necessary if the field was measured when the direction finder was properly oriented. For measuring E_n, this orientation corresponded to minimum response to E_n. By orienting the direction finder so as not to respond to E_n, it could not pick up and reradiate fields which would disturb the measurement of the field intensity. Such a procedure could not be used in the case of the spaced vertical and horizontal loop-antenna systems. For these cases the field was measured some distance to the side of the direction finder but the same distance from the transmitter. The assumption was then made that the attenuation of the wave was the same for this path as for the path to the direction finder.

When using a field-intensity meter employing a loop antenna, the electric field calibration made with plane waves does not hold when measurements are made in other than plane-wave fields, although a magnetic field calibration would. The field generated by the local transmitter is not the same as for a plane wave, the ground-reflected and surface waves being present as well as the direct wave. Under these conditions, to measure $E_{p,t}$ and E_n with the loop-antenna field-intensity meter,[a] the loop antenna of the field-intensity meter is oriented so as to respond to either H_n or $H_{p,t}$. The reading of the field-intensity meter, which will be in volts per meter, then refers to the related value of $E_{p,t}$ or E_n which would be present if the field measured were that of a plane wave in free space. This is what is meant, in this report, when H is measured by means of a loop-antenna field-intensity meter and designated as *volts per meter*; it is just the related value of E for a plane wave in free space. The reading of the field-intensity meter, though in volts per meter, will be a number proportional to H. The relation between E and H is given by the following equations:

$$H_n = -\frac{H_{ve}E_{p,t}}{E_{ve}\cos\psi}, \tag{66}$$

$$H_{p,t} = \frac{H_{vm}E_n\cos\psi}{E_{vm}}. \tag{67}$$

Here E_{ve} and H_{ve} are the values of the electric and magnetic radiation fields at a unit distance in free space in the equatorial plane of the electric dipole, while E_{vm} and H_{vm} are the corresponding values for a magnetic dipole.

Equation (67) is important because it shows that the loop antenna of the field-intensity meter should be oriented to respond to $H_{p,t}$ when measuring E_n, rather than for maximum reading of the meter as is done for plane waves in free space. The maximum reading corresponds to the amplitude of H_p which is often much larger than $H_{p,t}$, so that too large a value for E_n would be obtained. Since cos ψ is almost unity for these measurements it follows that $E_{p,t}$ equals the reading of the meter when oriented to H_n, and E_n equals the reading when oriented to respond to $H_{p,t}$. In most cases in this work the measurements of E_n were made by orienting the meter for the amplitude of H_p rather than $H_{p,t}$, since the correct procedure was not evolved until after most of the experiments were performed. This renders the measured values of E_n inaccurate for frequencies below about 7.5 mc where the wave tilt is appreciable. As already stated, the direction of this effect is such as to make the measured values of E_n larger than they actually are. Consequently the measured pickup factors, k, corresponding to E_n were too small and the calculated polarization errors also too small. This error was not made in the case of the United Air Lines [UAL] direction finder when used either with horizontal or vertical loop antennas.

After measuring the pickup factors of a direction finder the polarization errors were calculated using the method already given. In these calculations the response of the antenna system to $E_{p,t}$ involves an unknown phase angle Δ. An inspection of equation (27) for the balanced H-antenna system shows that the limits on the values of the maximum polarization error set by the unknown value of Δ are given by

$$\tan\epsilon = \frac{kE_n}{hE_{p,t}\cos\psi \pm kE_{p,x}}. \tag{68}$$

These limits can therefore be calculated. However, since $E_{p,r}$ varies as sin ψ, this unknown term will be small for low angles of elevation

and also for low values of k. For a direction finder having low polarization error, k is small so that the following approximate equation is valid:

$$\tan \epsilon = \frac{kE_n}{hE_{p,s} \cos \psi}. \qquad (69)$$

This is the equation used in practice for the calculation of polarization errors in this report, except for the loop-antenna direction finders. If k is so large that equation (69) is not a good approximation, the exact value of the polarization error is not needed because it will be very large. The equation can be used to indicate the superior or inferior performance of a direction finder.

In the results that follow, polarization errors are calculated for average ground conditions, that is, a conductivity $\sigma = 5 \times 10^{-14}$ emu and dielectric constent $K = 15$.

3.4.1 Test Conclusions

This experimentel technique evolved as the research progressed and included special means to overcome experimental difficulties encountered in the application of the method to measurement on particular direction-finder systems. (1) The stringent conditions for pure fields were relaxed for the case of Adcock direction finders by orienting the antenna system to the proper null position. (2) Remote control was used to control the radiator when generating perpendicularly polarized fields, while the direction finder itself was used as an indicator to tell when the radiator was exactly horizontal. (3) The influence of the direction finder on the measured field intensity was removed in many cases by properly orienting the direction finder. In other cases measuremente of field intensity were made off to one side of the direction finder. (4) Correction factors were computed and applied to allow measurement of electric field intensity by means of a field-intensity meter using a loop antenna. (5) To reduce errors caused by radiator parallax, a horizontal loop-antenna radiator was found necessary when generating the perpendicularly polarized wave field required in this work, while a vertical electric antenna was necessary for generating the wave polarized parallel to the plane of incidence. (6) Methods were developed for reducing the experimental errors caused by collector parallax for the case of particular direction finders.

The experimental technique finally evolved has important application to other methods of measuring polarization error. The errors caused by radiator and collector parallax will be present in the RCA and Barfield methods unless techniques similar to those used in this work are applied. Furthermore these methods encounter another difficulty, clarified by the work of this report, when measurements are made at angles of elevation below 20° to 30°. This is the error caused by the surface wave component of the field from the local transmitter and is not present in the NBS method.

Since the experimental technique evolved gradually, the measurements of polarization errors of the various direction finders were not all carried out with the same accuracy. In some cases E_n was incorrectly measured, giving measured polarization errors which were too small. This effect was unimportant above 7.5 mc. In other cases, radiator parallax was not avoided. Bearing in mind these limitations as to accuracy, Table 2 of approximate polarization errors compiled from these sections may still be used to draw certein important conclusions.

Table 2. Approximate polarization errors.

Direction finder	2.5 mc		5.0 mc		7.5 mc		10 mc		12.5 mc		15 mc	
	ϵ_v	ϵ_H	ϵ_v	ϵ_H	ϵ_v	ϵ_H	ϵ_v	ϵ_H	ϵ_v	ϵ_H	ϵ_v	ϵ_H
WE—CAA	41	45	15	37	8	30						
Experimental H antenna	21	20	13	23	12	36	4	20	11.7	51	5	11
SCR-551	22	12	12	14	11	20						
Navy DY	12	7	7	7	4	5	9	18	7.8	20	19	50
Collins CXAL	4.5		6.5		6.5		5.5		3.2		0.7	
Vertical loop antennas			16	9	10	7	1.2	1.1	1.0	1.1	1.1	1.6
Horizontal loop antennas	1.8	3	1.9	1	1.9	1	1.8	0.8	1.8	0.8	2	1.8

In Table 2 the values are given of two particularly significant "wave errors" which may be derived to represent the performance of a given direction finder: (1) the value of maximum bearing error for a downcoming wave incident at 45° with equal parallel and perpendicular components, ϵ_{45}, which is the "standard wave error" as defined by Barfield and (2) the value of maximum bearing error for a horizontally traveling wave also with equal parallel and perpendicular components, ϵ_h. The error ϵ_{45} includes the effect of the height of the direction finder antenna above ground and of the electrical properties of the ground whereas ϵ_h is independent of these effects.

Table 2 gives the horizontal wave error, ϵ_h, and the standard wave error, ϵ_{45}, in degrees. The value of ϵ_h is independent of the ground constants as already stated, while that of ϵ_{45} is for average ground having constants $K = 15$ and $\sigma = 5 \times 10^{-14}$ emu. The height of the various direction finders for which the values of ϵ_{45} are given is the height for which the direction finders were designed except in the case of the UAL antenna system. The height was taken as 10 feet for the vertical loop-antenna system as a practical value approaching optimum results. Two different heights were taken for the horizontal loop-antenna system; 50 feet over the band from 2.5 to 7.5 mc and 30 feet above 7.5 mc. In this way the whole frequency range was divided into two bands with approximately optimum antenna heights for each band. The values of ϵ_{45} for the experimental H antenna, the SCR-551 and the WE-CAA direction finders would have been lower for lower antenna heights. The complete data for each system are summed up in the graphs given in the particular section for that direction finder in the final report.[1, 19, 26]

The data given in Table 2 are shown in graphical form in Figures 12 and 13 in which ϵ_h and ϵ_{45} are respectively plotted as a function of frequency for each of the direction finders.

Inspection of Table 2 and Figures 12 and 13 shows that the polarization errors are in general much larger for those direction finders using open antenna elements than for those using loop antennas. This indicates that the loop antennas are inherently easier to balance

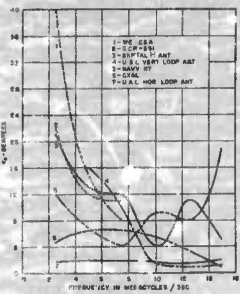

FIGURE 12. ϵ_h versus frequency for direction finders tested.

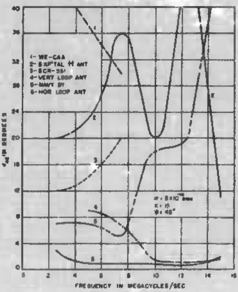

FIGURE 13. Maximum polarization error for equal plane wave components parallel and normal to plane of incidence.

and to shield properly so as to suppress unwanted pickup. The low impedance of the loop antennas is helpful on this score. This conclusion as to the relative superiority of the spaced loop-antenna direction finders is one of the principal results of this work.

Table 2 and Figures 12 and 13 also indicate that the direction finder having the least polarization errors of all those tested was the one using horizontal loop antennas. This superior performance was obtained without any critical adjustments of the antenna system. Measurements on this antenna system were free from radiator parallax errors and from errors of measurement of field intensity. Coupled with the low polarization error of this system is its low susceptibility to site errors. These two properties together would indicate a very promising system for those applications where sky waves only are being received and where the antenna may be placed approximately one-fourth wavelength above the ground.

Methods similar to those given in this report were used by RCA and International Telephone and Radio [IT & R] Laboratories to make polarization error measurements on the Bell Telephone Laboratories [BTL] buried U-antenna system at Holmdel, on the elevated, shielded U-antenna system of RCA, and on various IT & R systems at Great River, Long Island. It was found that the BTL buried U antenna had a performance similar to the loop-antenna systems described herein; the RCA antenna system had errors about the same as those of the electric antenna systems of this report, while IT & R reported an increase in accuracy for electric antenna systems achieved by using cathode followers to couple the antenna elements to the transmission line.

Critical consideration of the NBS method of measuring polarization errors as applied to the several direction finders shows that the method has several advantages. First may be mentioned the convenience and speed with which measurements are made since, in general, all measurements are made with the equipment near the ground and because waves polarized parallel and perpendicular respectively to the plane of incidence are used separately. This avoids the need for adjusting the phase of these two components as is necessary when both waves are used simultaneously. The fact that these waves are used separately also results in another important advantage, namely, that *maximum* polarization errors are measured.

This result is one of the principal results of the present research. Originally, Barfield defined the "standard wave error" to be the bearing error for the "standard wave" with such phase relation between the parallel and perpendicular wave components as to result in *maximum* error. However, the experimental technique employed by Barfield for many years did not meet the conditions required for maximum error; as a result the polarization errors measured were much too low. The publication of polarization errors measured by this method led to the general belief that polarization errors were quite small. Measurements by the NBS method and subsequently by that of RCA indicated much larger polarization errors for existing direction finders than had generally been believed to be the case.

In the Barfield technique maximum errors were not, in general, measured because the phase relation of the two components in the wave from the target transmitter was not adjusted for maximum error. Some months after first publication of the NBS method, an account appeared of recent attempts to modify the Barfield method so as to control the phase of the two components[15] but these had not yet been applied practically because of experimental difficulties. However, further measurements[16] of an H-antenna system,[17] in which allowance was made for the proper phase relation to give maximum error, showed errors two to ten times larger than those values given previously on the basis of measurements by the Barfield method. This result as to the extraordinarily large polarization errors of many present types of direction finders now agrees with that of the NBS and the RCA groups.

The large polarization errors found as a result of this work have refocused attention on the reduction of polarization errors. The NBS method has an important application to this problem of the reduction of polarization errors since it furnishes a figure of merit by which the progress of development work may be judged. After each change in design the pickup ratio of the antenna system may be measured

in order to determine the effect of the change on the polarization error. The technique is rapid and accurate.

The figure of merit proposed by the NBS and measured by the methods given in this report is the pickup ratio, h/k, of the direction finder. A practically equivalent figure of merit is the horizontal wave error, ϵ_0, as previously defined. The equation $\tan \epsilon_0 = k/h$ is the usual one given and is a means of translating the pickup ratio into an actual bearing error for an incident wave. The pickup ratio allows a direct comparison of polarization error for all direction finders following the same equation for polarization error and working on the same field components. The complete curve of polarization error versus angle of elevation of the incident wave should be used to compare the accuracy of antenna systems following different laws for polarization error. The pickup ratio is especially valuable for comparing the accuracy of direction finders because it is a fundamental d-f constant which is independent of the ground constants and the height of the antenna above the ground. In the case of buried U-antenna systems this constant is independent of the depth of the feeders below the ground even though the accuracy may be greater when the depth is increased, just as the accuracy of those systems above the ground, which are designed to suppress response to E_x, is increased by lowering the height of the antenna. Once the pickup ratio is measured, the polarization errors for any downcoming wave, such as the "standard wave," may be calculated for any antenna height or ground constants. This enables a study to be made of the optimum antenna height and ground constants for lowest polarization errors.

On the basis of such studies it was shown that the polarization error of a direction finder designed to utilize the E_y component of the incident wave and to suppress response to E_x components should be located over ground having the highest possible index of refraction. The choice of such a site requires methods of measuring the ground constants of proposed sites. A method was developed for this measurement which uses a field-intensity meter having a loop antenna. This method is easy to use and uncritical in its application. By measuring the ground constants at various points of the site, as indicated below, a test can be made of the subsurface homogeneity of the site and therefore its suitability from the standpoint of local site errors. Such methods and tests are becoming of greater importance because of the improved accuracy of newer types of direction finders. It is possible that polarization errors may eventually be reduced to the point where the bearing errors caused by the site may be of relatively greater importance. In this respect it may be important to use direction finders having inherently lower susceptibility to site errors caused by local reradiation. This research has shown that the spaced, horizontal loop-antenna direction finder should be relatively insensitive to reradiation errors because of the rapid attenuation by the ground of the horizontally polarized fields reradiated by surrounding objects.

4.4 DIRECTION-FINDER SITE PROBLEMS

The problems connected with d-f sites are numerous and complex. They may be classified, broadly, into two groups. The first group concerns the bearing errors caused by the site itself, that is, errors caused by deviation of the wave front or by reradiation. The second group concerns the effect of the site on (1) the direction finder or (2) the principal field at the direction finder. By (1) is meant the effect of the site in unbalancing the d-f antenna system, while by (2) is meant the effect of the site in suppressing undesired field components because of the interference between the direct and ground reflected waves. Furthermore, site errors can be classified as local or remote depending upon the distance of the source of the site error from the direction finder. Corresponding to this division into groups, the following discussion will take up the problems connected with the choice of a direction finder having the lowest susceptibility to site errors caused by reradiation, and those connected with the choice and testing of a suitable site. These site problems have recently assumed greater importance as a result of the improved accuracy of newer types of direction finders. It is possible that, excluding errors caused by lateral deviation, d-f accuracy will no longer be limited by

type, is believed to be relatively free from local site error.

Site Errors

Bearing errors caused by imperfections of the site have been classified as local or remote, although there is no sharp dividing line between these two groups. Remote site errors are usually caused by the character of the terrain. Large obstacles in the path of the wave, such as mountains, give rise to diffraction which results in a deviation of the wave front. Local site errors are caused by reradiation or reflection from nearby trees, wires, cliffs, etc. The random summation at the direction finder of all the reradiated waves gives a field which results in bearing errors. This distorting field can change rapidly with small changes of azimuth or frequency of the incoming wave.[13]

type, is believed to be relatively free from local site error.

This direction finder is designed to take a bearing on the E_n component of the incident field while ideally it should have no response to E_p. In general, reradiated E_r field components will be so severely attenuated by absorption in the ground that the reradiated E_n field intensity at the site will usually be very small. This is true even if the spaced horizontal loop-antenna direction finder is used at its optimum height $\lambda/4$ above ground. Figures 4 and 5 show that E_n will in general be equal to or greater than the perpendicular component in the incident wave, $E_{n,i}$. At this height, therefore, the effect of ground reflection on E_n will not be one of suppression. However, the field intensity at smaller heights will be decreased by the ground reflection so that reradiating objects at these smaller heights will have their effectiveness as

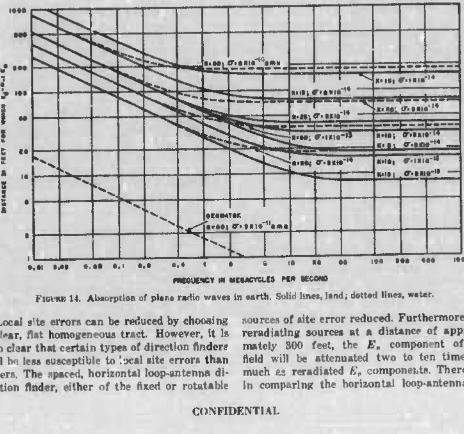

Figure 14. Absorption of plane radio waves in earth. Solid lines, land; dotted lines, water.

Local site errors can be reduced by choosing a clear, flat homogeneous tract. However, it is also clear that certain types of direction finders will be less susceptible to local site errors than others. The spaced, horizontal loop-antenna direction finder, either of the fixed or rotatable

sources of site error reduced. Furthermore, for reradiating sources at a distance of approximately 300 feet, the E_n component of the field will be attenuated two to ten times as much as reradiated E_p components. Therefore in comparing the horizontal loop-antenna di-

rection finder with those types designed to respond to the E_y component, it is clear that for equal field intensities in a downcoming wave, the total desired field will be approximately the same for the two types (used at optimum heights), while reradiated field intensities capable of causing bearing errors will be much less for the horizontal loop-antenna system. This comparison is made on the basis of equal amounts of energy reradiated from the disturbing source for the two cases.

6.3.2 Required Depth for Buried Cables

It is very convenient in many direction finders to be able to run cables, power lines, or telephone lines near the d-f antenna system. To avoid site errors caused by reradiation from such lines it is necessary to bury them an adequate distance below the surface of the ground. Figure 14 shows the absorption of plane radio waves in earth for various ground constants and for frequencies up to 1,000 mc. The distance in feet required for an attenuation of ten to one is shown on this diagram. It is interesting to note that the ultra-high frequencies are absorbed only slightly more than frequencies in the standard broadcast band in passing through media of average conductivity.

The field intensity at a depth Δ below the ground for a plane wave incident on the surface will be determined not only by the absorption of the wave but by the reflection at the surface. Previously, equations were given for the total field intensity at a height z above the surface. The field intensity at a depth Δ directly below the field point considered previously will be given by[d]

$$E_{p,z,t} = \frac{E_{p,d}}{n^2}(1+R_p)(\cos\psi)\, e^{(2\pi i/\lambda)(z\sin\psi + \Delta\sqrt{n^2-\cos^2\psi})}, \tag{142}$$

$$E_{p,x,t} = \frac{E_{p,d}}{n^2} \cdot (1+R_p)\sqrt{n^2-\cos^2\psi}\; e^{(2\pi i/\lambda)(z\sin\psi + \Delta\sqrt{n^2-\cos^2\psi})}, \tag{143}$$

$$E_{n,t} = E_{n,d}(1+R_n)\, e^{(2\pi i/\lambda)(z\sin\psi + \Delta\sqrt{n^2-\cos^2\psi})}. \tag{144}$$

[d] Equations (70) to (141) inclusive of the final report[1] are not given in this summary. The original equation numbers are retained here, however, for ease in referring to the original.

Here the subscript t indicates the transmitted wave. The term $2\pi z(\sin\psi)/\lambda$ in the above equations relates the phase of the transmitted wave to that of the incident wave at the height z above the ground. The attenuation factors in equations (142) to (144) may be determined from Figure 14 by identifying d with Δ and K with $K - \cos^2\psi$. This follows from the fact that the attenuation factor of a plane wave is given by e^{-Ad} where

$$\frac{2\pi n}{\lambda} = B + iA \tag{145}$$

$$A = \frac{2\pi}{\lambda}\sqrt{\frac{K}{\cos\alpha}}\sin\frac{\alpha}{2}, \tag{146}$$

$$\tan\alpha = \frac{X}{K}. \tag{147}$$

The absorption coefficient A may be determined for the case where d is expressed in feet simply by dividing the constant 2.303 by the distance in feet as given in Figure 14.

TABLE 8. Recommended depth to which underground lines must be buried to avoid reradiation errors.

[illegible]	Recommended depth for buried lines
[illegible]	
[illegible]	100 feet
[illegible]	50 feet
[illegible]	
[illegible]	20 feet
[illegible]	10 feet
[illegible]	
[illegible]	5 feet
[illegible]	2 feet
[illegible]	
[illegible]	8 inches

6.4 A METHOD OF MEASURING GROUND CONSTANTS

In choosing a site, it is desirable to have available quick and sensitive methods for testing its suitability without actually setting up the equipment and making bearing tests. For this purpose, visual observation of the flatness and freedom from reradiating objects of the proposed site is not sufficient, because the site must also be electrically homogeneous below the surface and must also have electrical constants falling within certain limits for best

results. A downcoming ionospheric wave may penetrate a considerable distance into the ground so that inhomogeneities located below the surface may reflect the waves strongly enough to cause bearing errors. To reduce this effect a site should be chosen over ground having as large an index of refraction as possible, such as a salt marsh. However, high conductivity and dielectric constant are also desirable from another standpoint. The site has a strong effect on the principal field at the direction finder because the total field is the vector sum of the incident wave and the ground reflected wave. This vector sum is termed the principal field in order to exclude fields generated by reradiating objects near the site. If the direction finder is designed to take a bearing through its response to E_r field components while its response to E_θ is suppressed, then it is desirable to locate the direction finder over a site having ground constants such as to suppress E_θ as much as possible. It has been shown in preceding sections that ground having a high index of refraction suppresses E_θ at points near the surface. Such ground also helps to screen buried lines and cable so that reradiation errors are reduced. All these considerations indicate that a quick method of testing a proposed site for subsurface electrical inhomogeneities and of measuring its electrical constants would be useful. NBS has considered a method of testing for inhomogeneities which uses a local oscillator and antenna near the ground. The variation over the site of the impedance reflected into the oscillator circuit by the ground reflection gives a test of the homogeneity of the site. Either the variations in the plate current of the oscillator or of the oscillator frequency could be used as an indication of homogeneity in these tests.

An alternative method of testing a site is to measure the ground conductivity and dielectric constant at various points of the proposed site. This method would not only determine the ground constants but also give a measure of the subsurface homogeneity of the site. Previous methods of measuring ground constants made use of electric antennas with their attendant difficulties. A new method of measuring ground constants by means of a standard field intensity set using a loop antenna, such as the RCA 308-A, will now be described[4] because of its application to the problem here considered.

In the previous discussion expressions for the fields generated by electric and magnetic dipoles near the ground were given for both vertical and horizontal dipoles. Equation (54) for the magnetic vector for the case of a radiating vertical magnetic dipole can be written as follows for the field intensity at the surface of the ground ($z = 0$); in this case $r_1 = r_2 = r$ and $\psi_1 = \psi_2 = \psi$ and we obtain

$$\mathbf{H}_p = H_{0m} \cos\psi \cdot [(1 + R_h) + (1 - R_h) f(P_m, B_m) e^{i\phi}] \cdot (\cos\psi\mathbf{k} + \sqrt{n^2 - \cos^2\psi}\,\mathbf{d}) \frac{e^{-ikr}}{r} \quad (148)$$

where $z = 0$ and $\mathbf{d} > \lambda$.

Equation (148) shows that at the surface of the ground both the space and surface wave components have the same polarization. The particular polarization of the magnetic vector and its forward tilt will depend on the ground constants and, if determined, give a means of measuring the ground constants. This is also true of the polarization and forward tilt of the electric vector $\mathbf{E}_p$ in the field of a radiating vertical *electric* dipole. However, in this case measurements of the electric vector would have to be made using electric dipole receiving antennas. Such measurements are difficult to make accurately because of disturbances to the electric field by the field-intensity meter and operating personnel. Therefore the new method based on measurements of $\mathbf{H}_p$ gives a preferable method of determining the ground constants.

Equation (148) may be written as follows:

$$\mathbf{H}_p = \mathbf{H}_{pd}\left(\mathbf{d} + \frac{\cos\psi}{\sqrt{n^2 - \cos^2\psi}}\mathbf{k}\right)e^{i\omega t}. \quad (149)$$

If we write

$$ae^{i\delta} = \frac{\cos\psi}{\sqrt{n^2 - \cos^2\psi}} = \frac{1}{\sqrt{K' - 1 - iX'}}, \quad (150)$$

where

$$K' = \frac{K}{\cos^2\psi}, \quad (151)$$

$$X' = \frac{X}{\cos^2\psi}, \quad (152)$$

and K is the dielectric constant of the ground

while $X = 1.797 \times 10^{11}\sigma_{emu}/f_{mc}$, then equation (149) becomes

$$H_p = H_{pd}\,[d \cos \omega t + k\,\alpha \cos(\omega t + \beta)]. \quad (153)$$

The above equation shows that the vector magnetic field from a vertical magnetic dipole rotates in an ellipse in the plane of incidence with its major axis tilted a few degrees above the horizontal. The magnetic vector reaches its maximum extension when $\omega t = -\delta$ and its minimum extension when $\omega t = -\delta + (\pi/2)$ where

$$\tan \delta = \frac{1}{2r}\left(\sqrt{1 + 4r^2} - 1\right), \quad (154)$$

$$\tau = \frac{\alpha^2 \sin 2\beta}{2\,(1 + \alpha^2 \cos 2\beta)}. \quad (155)$$

The measurable properties of the ellipses are θ, the tilt of the major axis above the horizontal, and r, the ratio of the minor to the major axis.

$$\tan \theta = \alpha(\cos \beta + \sin \beta \tan \delta), \quad (156)$$

$$r = \tan \delta \cot \theta. \quad (157)$$

Figure 15 shows θ and r as a function of X' for $K' = 5$, 10, 20, and 80.

The procedure for determining the ground constants from the above results is as follows. A small transmitter is used with a loop antenna which is set up with its axis in the plane of incidence at a distance greater than λ from the point at which the ground constants are to be determined and at a height such that an easily measurable field intensity is obtained at the receiving point. Figure 15 also shows a diagram of the experimental setup. Using a field-intensity meter with a loop antenna set up in such a manner that it can be rotated about an axis perpendicular to the plane of incidence but with the loop axis always lying in the plane of incidence, measurements are made of θ and r. The loop antenna of the field-intensity meter is to be placed as near the ground as possible because this procedure is based on equations derived for the case $z = 0$. Having measured θ and r, a corresponding set of values of K' and X' may be determined from the curves given in Figure 15. Finally K and σ are determined by

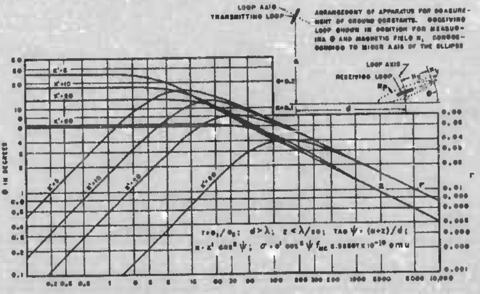

FIGURE 15. Polarization of vector magnetic field from vertical magnetic dipole.

means of equations (151) and (152) as follows:

$$K = K' \cos^2 \psi \tag{158}$$

$$\sigma = X' \cos^2 \psi \cdot f_{mc} \cdot 5.564 \times 10^{-14} \text{ emu.} \tag{159}$$

At very high frequencies, X' will be small and we may write

$$K' = \frac{1}{\sin^2 \theta}, \tag{160}$$

$$X' = \frac{2r}{\sin^2 \theta \tan \theta}, \tag{161}$$

where $X' << (K' - 1)$.

Equations (160) and (161) show that the dielectric constant K may be determined at very high frequencies simply by measuring θ while a determination of σ requires a measurement of both θ and r. This is also evident from Figure 15. At very low frequencies where X' is large, Figure 15 shows that the curves of θ and r are independent of the dielectric constant but either curve allows a determination of X' and therefore of σ. For this case

$$X' = \frac{1}{2 \tan^2 \theta} = \frac{1}{2r^2}, \tag{162}$$

where $X' << (K' - 1)$. Equation (152) also shows that the dielectric constant can not be measured at very low frequencies. However, this is no defect of the method since the dielectric constant has no appreciable effect on wave propagation at these low frequencies. As shown in Figure 15 the measurement of the ratio r of the minor to major axis of the polarization ellipse can be accurately made by means of a loop antenna and field-intensity meter provided only that the loop antenna is free from "antenna effect." This may be stated quantitatively as follows: the ratio of minimum to maximum reading in a *linearly* polarized magnetic field, such as is generated by a vertical electric antenna, must be much less than r.

Chapter 3

STUDY OF RADIO PULSE PROPAGATION

Pulses were transmitted from Puerto Rico and received at Holmdel, N. J., on a highly directional Muse system. Measurements indicated that direction finding on the first pulse of a pulse group gave significantly more accurate results than ordinary direction-finding methods, a fact of considerable value in long-range loran systems. Practically all the contractor's final report[1] is contained in this summary.

3.1 OBJECTIVE

This project[a] had as its objective the confirmation of certain ideas concerning the possibilities of long-distance short-wave direction finding and in particular the idea that there were times when measurements made on the first pulse of a pulse group would give a more accurate determination of the bearing of a station than would be obtained by ordinary d-f means.

Another object of the project was to obtain evidence as to what percentage of the time during which energy arrived over paths deviated from the great circle, energy also arrived over great-circle paths in sufficient amounts to operate a d-f system. For the period covered by the observations this condition existed for 80 per cent of the time.

3.2 DIRECTION FINDING SOURCES OF ERRORS

When energy is received over two or more paths, errors can be produced in certain types of short-wave direction finders even if the paths are all confined to the plane of the great circle passing through the transmitter and d-f locations.[2] These errors result from the fact that interference of the different components with one another produces instantaneous fields at each element of the antenna system, the phases and amplitudes of which are not determined solely by the wave direction and the geometry of the antenna system.

Furthermore, if one or more of the paths is deviated from the great circle, then practically all direction finders will give erroneous bearings. The extent of the errors and the percentage of time that they exist will depend upon the relative intensities of the components arriving over the various paths. Appreciable errors can be obtained even when the great-circle energy is greater than that arriving over the deviated paths.

Studies of short-wave radio transmission across the North Atlantic have disclosed two types of transmission phenomena which would produce such errors. During severe magnetic storms large amounts of energy have been observed arriving from the transmitter over paths which were widely deviated from the great-circle plane between the transmitter and receiver. At such times it has occasionally been observed that small amounts of energy arrive over a great circle path. At other times, during more or less normal transmission periods and on relatively low frequencies when energy arrives over several different paths, it has been observed that energy which has suffered several reflections at the ionosphere may be deviated appreciably from the great circle, whereas that which has suffered only a very few reflections will be deviated only very slightly if at all. Where d-f methods provide no opportunity of separately identifying the great circle and deviated path components, errors might therefore be anticipated during undisturbed as well as disturbed transmission periods.

By the use of short-pulse transmissions it is generally possible to separate the components transmitted over different paths on a time basis and accordingly to measure the direction of each path. When the different paths are all confined to the great-circle plane, direction finding on any of the pulses should therefore result in an improvement since those errors are eliminated which are caused by the interference of the various components with one another. However, if all of the paths are not con-

[a] Project C-35, Contract No. OEMsr-310, Western Electric Company.

fined to the great circle, and if the pulse upon which the measurements are made is chosen at random or because it is the strongest, errors in bearing would still be obtained. On the other hand, if the first pulse of a group is chosen it should generally give the most accurate bearings since it will have traveled over the most direct path. The work covered by Project C-35 was undertaken to verify this conclusion.

3.3 EXPERIMENTAL PROCEDURE

Pulses were transmitted from the University of Puerto Rico with equipment made available and maintained in operation through the efforts of G. W. Kenrick. A small rhombic antenna directed towards New York City was used. The bearing of the transmitter from Holmdel is 160° measured clockwise from true north and 1,581 miles distant. The transmitted pulses were 100 microseconds long and had a peak power of about 1 kw. The recurrence rate was 60 per second except for some of the preliminary experiments when rates of 20 and 30 per second were used.

Measurements on the direction of arrival of the individual pulses were made with the Holmdel Musa receiving equipment in accordance with a procedure described in a previously published paper.[3] As pointed out in that paper, two sets of antennas with different axes of orientation can be used in connection with the Musa equipment to determine the actual direction of arrival of the waves in space. For these experiments only two antennas of each set were used instead of the usual six and the phase shifters were adjusted for cancellation instead of addition. This permitted the use of two widely spaced antennas of each set thereby giving greater accuracy in the bearings. Check measurements were made occasionally with closer antenna spacings (adjacent antennas) in order to avoid ambiguous results. The band width of the receiving equipment was sufficient to pass the pulses as transmitted without appreciable alteration in their shape.

Pulses were transmitted on three different frequencies; 17,310, 7,175, and 6,425 kc. During the first month pulses were transmitted on 17,310 kc during the daytime and on 6,425 kc during the evening and nighttime hours. Observations were made during two hours in the morning and two hours in the evening for three days a week. During the last two months the schedule was changed. Pulses were transmitted continuously on 17,310 kc for the first half of each week and on the lower frequency during the second half of each week, thus permitting observations to be made on either frequency during any desired hour of the day or night. During these last two months special attention was paid to the transmission conditions existing during the sunset period since it had been observed that rather wide deviations in the direction of arrival occurred at that time.

The 7,175-kc frequency was substituted for 6,425 kc during the last few weeks of the tests because the interference on the latter frequency became so severe that reliable measurements could not be obtained.

3.4 RESULTS

Measurements were made between January 12, and March 28, 1942, inclusive, on a total of 185 pulse groups on 17,310 kc and on 87 pulse groups on 6,425 and 7,175 kc, the data taken on these last two frequencies being grouped together. Some of the pulse groups consisted of only one or two distinct pulses while others consisted of five or six or more pulses, some of which overlapped to such an extent that the individual pulses were indistinguishable. Of the 185 groups measured on 17,310 kc only 5, or 2.7 per cent, contained pulses which arrived over paths deviated by more than 2° from the great-circle path to the transmitter and the maximum deviation was only 3°. Of the 87 groups measured on 6,425 and 7,175 kc, 35 or 40 per cent contained pulses which arrived by paths deviated by more than 2° from the great-circle path. The greatest deviation measured was 12.5°. These results are shown graphically in Figures 1A and 1B where the deviations from the true bearing are plotted as abscissas and the number of pulse groups containing pulses with a given deviation are plotted as ordinates. Since the experimental error varied from 1.5° to 2°, depending upon the frequency used, observed deviations of 2° or less are not considered as significant and accordingly are not shown on these graphs.

Four of the five pulse groups on 17,310 kc and twenty-nine of the thirty-five on 6,425 and

7,175 kc which were observed to contain pulses which arrived over deviated paths also contained pulses which arrived earlier and over paths deviated by not more than 2°. In some of these cases the energy arriving over the deviated paths was appreciable so that bearing

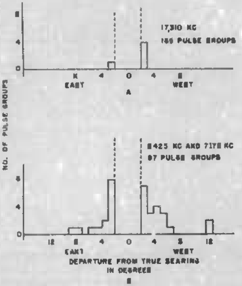

FIGURE 1. Percentage of pulse groups containing pulses deviated from great circle.

measurements on the first pulse of the group would have given significantly more accurate bearings than ordinary d-f measurements. In the remaining one of the five 17,310-kc groups and in the remaining six of the thirty-five lower frequency groups, no earlier true bearing pulses were observed within the time limit of thirty minutes allowed for the measurements to be considered as including only a single pulse group. It is entirely possible that even in these few cases there were less deviated paths that would have become evident had higher-powered pulses been available.

DISCUSSION

Aside from the improvement to be gained by direction finding on pulses in general, it was found that, under similar conditions as to path length and location, direction finding on the initial pulse on frequencies around 17 mc would result in only a very slight improvement in accuracy, but on the lower frequencies the improvement would at times be appreciable.

This lack of expected improvement on the high frequency results from the fact that the transmission on these frequencies is generally confined very closely to the great-circle plane. This is in accord with previous experience that, in general, the higher frequencies are better than the lower frequencies for d-f purposes. This seems to be true, not only during normal undisturbed days, but also during magnetic storms, the reason probably being that h-f transmission takes place by lower angle paths with fewer reflections at the ionosphere so that it is less adversely affected by variations in that medium.

During the period over which these experiments were conducted there was only one short severe magnetic storm and no measurements were taken during the height of that storm. Observations made during the following days and during other slightly disturbed periods indicated that for this particular path the only effects were a decrease in field strengths and a very slight increase in the number and extent of the deviations observed on the lower frequencies. This lack of a pronounced magnetic storm effect is not inconsistent with previous observations, for it has been observed that radio paths which pass near the magnetic poles are in general much more severely affected by magnetic disturbances than those paths which are distant from the poles. If more conclusive evidence of the improvement to be expected during disturbed periods by initial pulse measurements is desired, it is believed that pulse transmissions over a path much nearer the magnetic pole than the one used for these experiments will have to be studied.

In the light of past experience with continuous-wave transmission over the North Atlantic path and present experience with the pulse transmissions from Puerto Rico, it is felt that it can safely be predicted that direction finding on the first pulse will give a significant improvement in accuracy for a large percentage of the time during magnetic storms for transmission paths near the magnetic poles.

Those engaged in short-wave d-f research

recognize that one of the most severe conditions under which direction finders must operate occurs when the d-f location is just outside the ground-wave range of the transmitter and still so close to it that the ionospheric waves arrive at very high angles of incidence. Under such conditions the sensitivity of most direction finders to the desired polarization is very low so that any errors caused by spurious pickup are greatly accentuated. Furthermore any slight irregularities in the ionosphere can cause the path of the waves to be deviated considerably from the great-circle plane. If the ground-wave range were extended considerably for such cases by increasing the peak power of the transmitted signal, as can be done using the modern pulse technique, and if bearing measurements are taken on the first or ground-wave pulse, considerable improvement in accuracy would be expected. To test these conclusions would require a high-powered pulse transmitter located relatively close to the receiver location. The experiments discussed above do not apply to this case at all since the distance was entirely too great for the ground wave to be effective with the power used.

Chapter 4

ULTRA-HIGH-FREQUENCY DIRECTION-FINDING STUDY

Study of theoretical and practical aspects of wide-band directive antennas for direction-finding [d-f] use in 150- to 800-mc region led to development of two antennas—a corner reflector arrangement and a flat reflector, in each case with the array before the reflector. Proper phasing of array elements before the flat reflector rotated the directivity in azimuth. Design of a transformer for converting a balanced to an unbalanced system, use of new methods for evaluating polarization errors and for measuring electrical characteristics of the ground, and studies of the impedance characteristics of cylindrical dipoles of large transverse dimensions, formed a part of this study. The project[a] is reported rather fully here, the chief abridgement of the contractor's final report[1] lying in the omission of certain photographs of the equipment and certain charts that resembled closely those reproduced herein.

4.1 INTRODUCTION

Prior to World War II, d-f systems operating in the u-h-f region were generally of the elevated H, fixed or rotatable, Adcock type. Their properties had been extensively studied and were well known. On the other hand, later work with certain types of arrays and their application to radar and closely related fields indicated that improved systems having considerably higher gain and broad-bond response, particularly where portability was an important factor, could be devised for d-f use.

The studies in this project, therefore, consisted of the design of reflectors and arrays of the corner- and flat-reflector types; of means for rotating the directivity of the flat reflector by phase adjustment of the array elements; of the use of cathode-ray oscilloscopes for visual indication of bearing including electrical circuits for obtaining CRO patterns easily interpreted; and, finally, some comparative studies of a differentially connected V array and the conventional elevated H Adcock.

[a] Project No. 13.1-82, Contract OEMsr-1009, Radio Corporation of America.

4.2 RESEARCH FACILITIES

The site selected for this study is located on flat farm land near Medford, New Jersey, in an area known geologically as the Middle Marle Beds. The land is chiefly soil with small proportions of sand and clay and is known to be homogeneous to a considerable depth. A 10x12-foot building was erected to house the equipment and a 90-foot pole and rigging was installed for making the polarization error measurements. So far as possible the building was constructed of nonconducting materials. Wood and masonite were used as the basic materials and, with the exception of removable metallic window screens, metallic reflecting surfaces were kept to a minimum. The pole for supporting the polarization test transmitters was equipped with a removable carriage raised or lowered by means of a windlass. Wooden dowels were used in place of nails or bolts in all the structure above a fixed platform surrounding the pole and located at the same elevation as the roof of the test house to permit measurements at horizontal incidence of the array located on the roof.

It was found later that complete elimination of metallic objects in the construction of the equipment on the pole was not necessary at the frequencies used, and that metal could have been employed in limited amounts in the windlass and possibly in the pulleys. The effects of the metallic window screens were negligible since the screens were not in the line of the direct or ground-reflected waves at the transmitter. Presence or positions of persons or objects in the test house had almost no influence on bearings from the two types of arrays tested.

Power obtained from lines 600 feet away came to the pole in a shielded conduit buried to a depth of 18 inches and to a depth of two feet between pole and house. A six-conductor line

between house and pole (to provide meter outlets at the house for circuits located at the pole) was buried to a depth of two feet in a trench containing the telephone circuit. The

FIGURE 1. Elevation view of test site. Pole and measurement house were [illegible] ft apart.

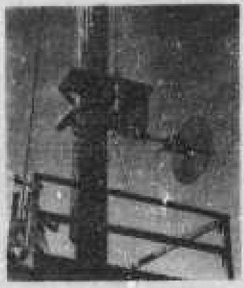

FIGURE 2. Carriage employed for hoisting transmitter for making polarization measurements.

discontinuity in the ground characteristics caused by the buried cables was not serious as subsequent site error measurements proved.

4.2.1 Receiving Equipment

The receiver was an experimental model SCR-616 supplied by the Eatontown Signal Corps Laboratory. It covered the bands 150–300 and 300–600 mc. It consisted of a superheterodyne receiver with one r-f stage on the lower frequency band and no preselector on the higher band. The input was designed to operate from a 95-ohm balanced line. Satisfactory results were obtained in most cases by operating the receiver from an unbalanced line.

The receiver had good performance characteristics for the service required. It was fitted with an a-f injection oscillator to modulate the intermediate frequency to produce an audio component when receiving pure c-w. The receiver sensitivity, expressed in microvolts to a 95-ohm dummy antenna, modulated 30 per cent at 400 cycles, required to produce a change in output voltage of two to one with carrier on and with modulation on and off respectively, was approximately 5 microvolts on the lower frequency band and 15 microvolts on the higher band.

4.2.2 Measuring Equipment

A slotted transmission line was employed for impedance measurements and for a wide range of other necessary measurements. A microammeter equipped with a tilted mirror was found convenient for making observations when it was mounted on the elevated car iage and read with the aid of high-powered prism binoculars from the ground. A General Radio power oscillator covering the range of 150–600 mc furnished signals.

4.3 V-1 ARRAY (1 DIPOLE PER SCREEN)

The V type of antenna system consists of two similar linear cophasal broadside arrays, each placed in front of one side of an angled reflector. The two arrays have mirror-image response patterns, each nearly symmetrical, excepting that one is rotated in azimuth with reference to the other by an angle equal to the angular displacement of its reflector from the plane of the other. The direction of maximum response of each array is normal to the corresponding reflector. See Figures 3A and B.

It may be observed that the patterns of the two arrays can be made to intersect at some desired point, depending on the angle of the reflector. This intersection represents equal response of the two arrays, and may be used as a bearing indication if the amplitudes are suit-

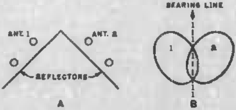

FIGURE 3. A shows typical V array. B shows typical response patterns of individual antennas.

ably compared in associated indicating equipment. A number of indicating methods are available and are discussed later.

Two collector systems of this general type were studied, the first consisting of one dipole before each reflector (V-1), and the second having two dipoles in front of each reflector (V-2). Larger numbers of cophasal antennas per screen are possible, but were not studied because the resulting size in the low-frequency band was considered excessive in view of portability requirements. On the higher-frequency band, twice the number of elements may be used without exceeding the size of the low-frequency array.

A large portion of the experimental work in this project was done on the V array having one dipole per reflector, and while this array has the least favorable performance of all considered, most of the information obtained was useful in carrying out the examination of the other arrays.

4.8.1 Reflectors

The first problem presented was the determination of reflector size and mesh. The experience of other groups engaged in antenna research indicated that a reflector approximates a perfect plane conductor of infinite extent if the dimensions are such as to exceed by one-eighth wavelength in all directions the maximum dimensions of the array with which it is used at the lowest frequency of operation, providing that the smallest dimension of the array is at least one-half wavelength ($\lambda/2$) at this frequency. A maximum spacing of $\lambda/20$ between members making up the screen at the highest frequency to be used was indicated, with the length of the elements oriented along the direction of desired polarization. The present study verified the adequacy of these limits. An increase in the screen size above this figure resulted in small performance change. Substitution of high-conductivity fine mesh screen also resulted in no material improvement in performance.

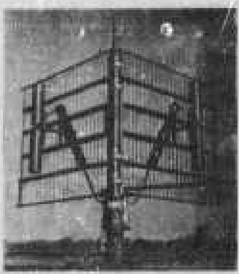

FIGURE 4. V-1 array in front of screen. Coupling transformers are behind screen at angle to antennas.

The screens used were fabricated of 3/16-inch stainless steel tubing, spaced two inches apart, and made in sections so that the overall size was easily adjustable. The members supporting the dipole assemblies were designed to permit adjustment of the spacing between dipoles, and the spacing to the screen. The assembly was copper plated and protected by a coating of enamel. See Figure 5 for dimensional drawing.

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4.5.8 Dipole Dimensions and Impedance Characteristics

The determination of proper dipole dimensions required considerable attention. The parameters chiefly affected by the dipole dimensions are the radiation impedance and its resistive and reactive components. The efficiency of energy transfer from the dipoles to the utilization circuits depends on the impedance match through the system (aside from line losses); it was therefore necessary to establish some criteria to guide the work toward obtaining a desirable characteristic.

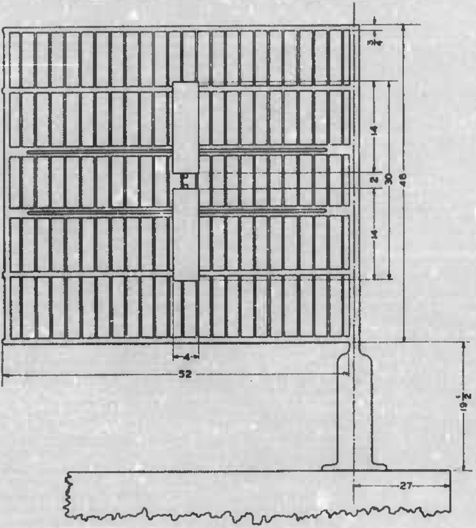


FIGURE 5. Essential physical dimensions of dipole and its reflector.

The input circuits of receiving equipments in the range of frequencies covered, 150-600 mc, differ markedly from lower frequency equipment in that power is generally consumed in the former due to an input impedance characteristic which may reach low values. This is due primarily to finite electron transit time effects which depend on the size of the first amplifier tube, its geometry, and electrode voltages. The input conductance varies directly with the square of the frequency. As a consequence receiving equipments may have widely different input impedance characteristics, and the input impedance of a receiver will generally show a large variation over a two-to-one frequency range. This impedance is primarily resistive since the circuit is usually tuned. With an antenna and receiver whose impedances are different functions of frequency, a matched condition can not be realized with a fixed transmission line over a wide band of frequencies. In most cases the condition for best signal-to-noise ratio corresponds to the condition of maximum power transfer. This latter condition requires that at any point in the system the impedance in one direction must be the complex conjugate of the impedance in the other direction.

In view of these facts it is considered impossible to set up absolute criteria for the impedance characteristics of an antenna array without a complete knowledge of the equipment with which it is to be used. An alternative which is thought to be satisfactory is to approximate a uniform resistive impedance through the range. The mismatch between this and a suitable transmission line should not be too severe. The input circuits of receivers appear to offer a greater degree of flexibility for applying means to manipulate impedance characteristics, and an attack of the problem in this direction should more readily yield the desired results. It is not unlikely that incomplete information in the hands of receiver designers on wide-band antenna impedance characteristics is one of the chief reasons why such large variation is encountered in the input circuits of receivers. In comparison two of the systems developed in this project, the V-2 and flat arrays, show much smaller variations, while the variation of the V-1 is of the same order as a representative receiver covering a similar frequency range.

The preliminary design work on the corner array envisioned the use of dipoles mounted by metallic tubes supporting each half, the two tubes forming in effect a parallel wire transmission line shorted at the reflector surface. Electrically this represents an almost pure reactance in shunt with the radiation impedance of the dipole. This shunt reactance was expected to cancel partially the radiation reactance of the dipole; the length and characteristic impedance were chosen so as to accomplish this.

Preliminary measurements showed large discrepancies between actual and expected results, due largely to insufficient information on the characteristics of cylindrical dipoles of large transverse dimensions, the design work having been based on prolate spheroidal dipoles. Also, the effect of the reflector was not fully accounted for. The results of subsequent theoretical investigations on these points are given below.

As a result of information obtained in these measurements, it was decided to change the design to a single-support, insulated dipole, using twin coaxial cables for interconnection. This eliminates the shunt reactance of the double support, and replaces it with the much higher reactance of a single insulated support. This balanced configuration was expected to be less susceptible to response from fields of undesired polarization.

Dipole Considerations

The frequency range which a dipole must cover restricts the choice of its length; this should be a half wavelength near the center of the band. The other constants which may be adjusted are the ratio of diameter to length, and the spacing before the reflector. A large ratio of diameter to length gives a low antenna characteristic impedance and resulting low Q. The limitations in this respect are chiefly mechanical, and depend on the portability required in the equipment. Weight and mechanical strength and rigidity are more favorable in the smaller diameter dipoles, and they are more easily supported, particularly in the balanced type of structure requiring members insulated

from the support. The only practical limitations on the spacing before the screen are the effect on the impedance variation and the change in the response patterns. With a single dipole before a screen, the response pattern begins to break up into two lobes as a quarter-wavelength spacing is exceeded. This is most pronounced at the high-frequency end of the range, where the fractional wavelength spacing is the greatest. When this condition is encountered, the antenna gain drops, and internal screen angles much less than 90° are required to obtain satisfactory overlapping of lobes. With arrays of more than one element per screen, this limitation is less serious, since the gain is higher and the response patterns are sharper.

Impedance Considerations

The effect on the impedance due to the spacing from a reflector may be examined most readily by replacing the screen by the negative image of the dipole. The mutual radiation impedance between the two dipoles modifies the impedance of the original dipole. The mutual reactance may be either positive or negative, and the mutual resistance may also be of either

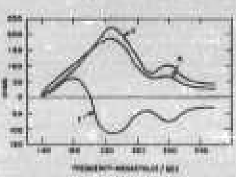


FIGURE 6. Impedance characteristics of V-1 array. Antenna-to-reflector spacing is [illegible] cm [illegible]

sign, depending on the spacing in terms of wavelength. The resistance decreases when the mutual resistance is positive, and increases when it is negative. When the self and mutual reactances are of the same sign, the reactance decreases, while if of opposite sign, an increase takes place. Since these mutual effects are dependent on the proximity of the dipole and its image, the impedance variations through a wide frequency range are generally greater when the spacing is small. The impedance characteristics, as measured at the dipole terminals, of the final V-1 array are shown in Figure 6.

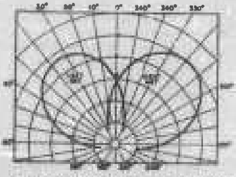


FIGURE 7. V-1 array, relative response in azimuth, 0° elevation, [illegible] mc.

4.3.3 Directivity in Azimuth

The directivity of this array in azimuth results from the use of the reflector, since the pattern in the equatorial plane of a dipole in free space is circular. As the frequency is increased so that the spacing from the reflector exceeds $\lambda/4$, the pattern begins to break up

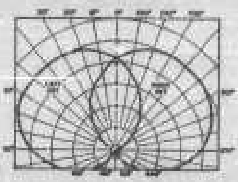


FIGURE 8. V-1 array, relative response in azimuth, 0° elevation, 300 mc.

into two lobes, with a minimum normal to the reflector. This minimum reaches zero at a spacing of $\lambda/2$. As a result, the spacing may not

appreciably exceed $\lambda/4$ at the high-frequency limit. The patterns exhibit a broadening with increasing frequency because of this phenomenon, but are seen to be usable through the two-to-one frequency range.

The general shape of azimuthal curves at 150 and 300 mc for elevations from 0° to 30° at 10° intervals changes somewhat through this range, but not sufficiently to affect the performance adversely (see Figures 7 and 8). The change in relative size of the patterns of the two antennas is due to the unsymmetrical effect of the horizontal electric field component lying in the plane of propagation; the shift of the lobe intersection from 0° azimuth is, therefore, a polarization effect.

4.2.4 Determination of Lobe Intersection

In d-f systems using amplitude comparison, the question arises as to the best point to use in intersecting the lobes, where such choice exists. The lobe intersection in the V array is determined by the angular position of the two reflectors, and can be adjusted to any desired point. It is, therefore, useful for this as well as other similar systems to consider a number of factors which enter into the selection of the intersection point, and if possible, to define an optimum point.

Consider a simple array such as the corner type with one element per reflector. If an idealized point antenna is placed before a screen, the relative response as a function of the azimuth angle measured from the normal to the reflector is given by

$$r = \sin\left(\frac{2\pi s}{\lambda}\cos\phi\right)$$

Here r is the relative response, ϕ the azimuth angle, s the spacing from the reflector to the point antenna, and λ the wavelength. This function is plotted in Figure 9 with s taken equal to $\lambda/4$. If a receiving equipment having an ideal noise characteristic, that is, one producing no internally generated noise, is thought of as being used with this antenna, then an output of any desired magnitude may be obtained from a signal arriving from any direction where the antenna response is not zero, assuming unlimited amplification to be available. In an actual receiving system the inherent noise limits the useful amplification. If an actual receiver is considered operating with the antenna, a certain amount of noise will be present in the output. If a signal now arrives at the antenna, the output of the receiver will be proportional to the field intensity, and to the relative response r of the antenna at the azimuth of wave arrival. (The gain of the antenna need not be considered at this point since its effect is merely to change the factor of proportionality.) For a fixed field intensity of the signal, therefore, the signal-to-noise ratio of the output is proportional to r, and from this standpoint best operation is had when r is a maximum.

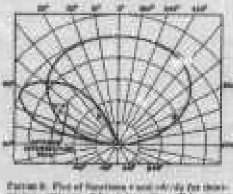


FIGURE 9. Plot of functions r and $dr/d\phi$ for determining optimum lobe intersection, V-1 array, single-point antenna $\lambda/4$ before screen.

The operation of indicating systems used in conjunction with switched lobe antennas usually depends on the difference in output when the signal is sampled successively on the two antennas. For example, a differential rectifier might be used to actuate a zero-center microammeter, or to control the input circuits of a servoamplifier for automatic tracking.

Sensitivity Factors

The sensitivity of such a system depends on the magnitude of the difference response for an increment of azimuth angle in the vicinity of the equisignal point, that is, the intersection point of the two lobes. The magnitude of the difference in turn depends upon two factors—the first of which is the slope of the antenna

response curve at the operating point; this quantity might be appropriately termed "differential sensitivity." The second is the scale factor, which depends on the amplification and signal intensity. If we assume that the maximum available linear amplification is used, then for a fixed signal intensity the differential output is proportional to $dr/d\phi$. As previously indicated, the signal-to-noise ratio is proportional to r. The conditions of maximum signal-to-noise ratio and differential sensitivity cannot be satisfied simultaneously, since $dr/d\phi$ is zero when r is a maximum. If equal importance is assigned to r and $dr/d\phi$ the maximum value of the product may be defined as the optimum intersection point. That is,

$$\frac{d}{d\phi}\left(r\frac{dr}{d\phi}\right) = 0.$$

The product $rdr/d\phi$ may be most conveniently maximized graphically, especially when experimental response curves are used.

For the case of a single-point antenna placed a distance $\lambda/4$ before a screen the functions

$$r = \sin\left(\frac{2\pi s}{\lambda}\cos\phi\right)$$

and

$$r\frac{dr}{d\phi} = -\frac{\pi s}{\lambda}\sin\phi\sin 2\left(\frac{2\pi s}{\lambda}\cos\phi\right)$$

are plotted in Figure 9. The maximum value of the latter is seen to occur at ϕ of approximately 60°.

For the case of two-point antennas placed before a screen, the response pattern is given by

$$r = \cos\left(\frac{2\pi d}{\lambda}\sin\phi\right)\sin\left(\frac{2\pi s}{\lambda}\cos\phi\right)$$

and

$$r\frac{dr}{d\phi} = -\frac{\pi s}{\lambda}\sin\phi\cos^2\left(\frac{2\pi d}{\lambda}\sin\phi\right)\sin 2\left(\frac{2\pi s}{\lambda}\cos\phi\right)$$
$$-\frac{\pi d}{\lambda}\cos\phi\sin^2\left(\frac{2\pi s}{\lambda}\cos\phi\right)\sin 2\left(\frac{2\pi d}{\lambda}\sin\phi\right).$$

Here the quantity d represents half the distance between the two dipoles. These two functions are plotted for $d = s = \lambda/4$ in Figure 10. The maximum of the second occurs at ϕ of approximately 28°.

This figure of merit fails under extreme conditions of signal intensity. If the signal is on the threshold of noise, r becomes more important than its derivative, while at the other extreme, where the receiving system is overloaded, operation at lower values of r is indicated. The first of these extremes is more likely to occur in practice; however, deviation from the above criteria should be based on statistical data obtained in actual use in the field. The data should include the noise charecteristics of the receiving equipment and the field intensities encountered. The speed of response of the indicating circuits, or the minimum time in

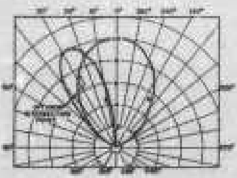


FIGURE 10. Data similar to that given in Figure 9 except for two-point antenna. In both illustrations, $rdr/d\phi$ is product of azimuthal response and rate of change of this response with respect to azimuthal angle.

which integration is substantially complete, coupled with the signal-to-noise ratio data should indicate the direction and extent of the departure required from the above criteria.

In the two cases considered, the $rdr/d\phi$ curves are fairly broad near the maxima; for one doublet, the width of the curve is 20° at 90 per cent of maximum, while for two doublets the width is 16°. This width affords some latitude of choice without departing far from the optimum. When an array is used over a two-to-one frequency renge, the shape of the response pattern changes with frequency. The intersection point can be selected at the mean frequency, and performance will generally be satisfactory throughout the range.

Relative Response in Elevation

Because of interference effects between the direct and reflected waves at the receiving antenna, the direct measurement of the directive pattern in elevation holds only for a specified antenna height above ground and fixed ground constants. For this reason an indirect method was adopted for the measurement, and the patterns so obtained are assumed to hold for all the arrays studied in this project

The method of measurement consists of determining the response in azimuth of one reflector to horizontally polarized waves, with the receiving dipole oriented horizontally. Since the reflector used is very nearly square, this procedure, in effect, is equivalent to turning the entire receiving and transmitting system 90° about a horizontal line connecting the transmitting and receiving points. The original vertical polarization now corresponds to horizontal, and the angle of elevation corresponds

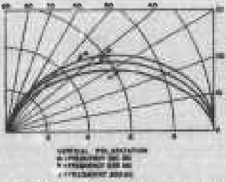


Figure 11. V-1 and V-2 relative response in elevation, [illegible], in half-space above the earth in absence of ground reflections.

to azimuth. Ground reflections are thus constant, since all measurements are made at 0° elevation, and may be neglected. The measurement was made out to ±90° from the normal to the screen; the two halves showed slight dissymmetry, and the average was taken. The response diagrams for three frequencies, 150, 225, and 300 mc are given in Figure 11, and may be taken to represent the relative response in the half space above the earth, in the absence of ground reflections. It appears to serve no purpose to give specific diagrams including the effect of ground reflections, since these would vary widely depending on the height, frequency, and ground constants.

Polarization Errors

One of the major problems encountered in the course of this project, and one which arises in most research connected with the study of collector systems for direction finders, is the investigation of polarization errors. Although most of the important characteristics of such systems can be readily determined theoretically or experimentally, this is not true in the case of polarization errors. The theoretical prediction is generally not possible, except in the case of certain elementary collectors such as balanced shielded loops, where the response to fields of any polarization is known. In the case of some other antennas of simple geometrical configuration, the shape of the response pattern to fields of various polarizations may be assumed to a good degree of accuracy; the errors may be predicted if the scale factors between them are known. The latter cannot be predicted theoretically, and therefore are measured; the complete performance can then be stated in terms of the theoretical assumptions and the measured values of these parameters.

The difficulties underlying the evaluation of polarization errors are due basically to the lack of an adequate and readily measured standard of performance. Until recently, the standard wave error of Barfield[a] was widely used. This is defined as the error of a system when obtaining a bearing on a wave arriving at an angle of elevation of 45°, and having equal components polarized in, and perpendicular to, the plane of incidence, with the two components so phased as to produce the maximum error. This standard of performance is open to two objections. The first is that it is defined without consideration of the effect of the ground in modifying the wave arriving at the collector. Although the omission is justifiable in the case of collectors located near the ground in terms of wavelength, in the case of elevated systems the difference in reinforcement or cancellation of the parallel and perpendicular field compo-

nents caused by the ground-reflected wave may give rise to a resultont that differs widely from the condition of the downcoming wave. Hence, large variations of the standard wave error of a system may be observed, depending on the elevation of the receiving antenna above ground and the electrical characteristics of the ground. The second objection is that a knowledge of the standard wave error of a system in itself is generally not sufficient to determine its performance under other conditions of wave arrival. The additional information necessary for this determination is the law of error which the system follows. The general law which a system obeys is, of course, known from the theory of its operation; to reduce it to a quantitative form usable in extrapolating errors, other data would appear to be necessary. This same objection put in another form applies when systems following different laws are intercompared. Evidently one system may have a very large error at an angle of elevation of 45° (for example, if the vertical response is a minimum at this angle and the horizontal response is large), and have low errors at other elevations. A second system may respond in an opposite manner, and have low errors at 45° and high elsewhere. Comparison of these two on the basis of the standard wave error would appear favorable to the second, while the first may actually be superior at most other elevations. Therefore it is seen that two arrays having the same standard wave error may have considerably different performance under other conditions.

To some extent the situation has been clarified by the work of the National Bureau of Standards [NBS] as summarized in the final report on Project C-18.[8] The report is condensed in Chapter 1 of this volume. The methods and criteria developed by NBS for the evaluation of collector performance with regard to polarization errors overcome these objections to a certain extent and should be applicable, in theory at least, to most collector systems. These methods specify performance in terms of certain parameters of the system analogous to effective heights, measured at ground level, and independent of the ground constants. A knowledge of the law of error of the system enables the complete performance to be stated. It was thought desirable to apply these methods in the present project, subject to verification of the results by other methods, principally, the direct measurement of the errors. Unfortunately, the attempt to adapt these methods in the present case did not result in any marked degree of success.

The NBS Method

Essentially, the NBS method is based on the statement of the response of the antenna system to an arriving field in terms of the desired response of the true antenna elements and the undesired response due to extraneous elements such as feeders, etc. The response of the antenna is analyzed on the basis of three resolved field components, with a directivity function associated with each, and a parameter analogous to effective height, called the pickup factor, also associated with each. These latter relate the voltages induced by a component to the intensity of the component producing it, and therefore have the dimensions of effective height. The response of the feeders is simliarly stated for the three field components. The equations may be written as follows:

$$V_{\text{ant}} = hEF(\phi,\psi) \qquad (1)$$

$$V_{\text{feeders}} = kEf(\phi,\psi) \qquad (2)$$

Here V is the voltage induced in the element indicated in the subscript, h and k the pickup factors, E the electric field intensity (for simplicity the magnetic field components will not be considered), and F and f the directivity functions, dependent on the azimuth ϕ, and the angle of elevation ψ. The field terms on the right hand side of the two equations are resolved into three components, and the other two factors are likewise resolved to correspond.

The resolution of the field at the collector is indicated in Figure 12, where the two primary components E_n and E_p, shown at A, are respectively perpendicular to and in the plane of incidence.

The parallel component is further resolved at B into a vertical component E_{pv}, and a horizontal component E_{ph}; the direction of propagation associated with each vector is indicated

in the figure. Following this resolution, equations (1) and (2) can be written

$$V_{\text{ant}} = h_z E_{p,z} F_z(\phi,\psi) + h_x E_{p,x} F_x(\phi,\psi) + h_y E_n F_y(\phi,\psi) \quad (3)$$

$$V_{\text{feeders}} = k_z E_{p,z} f_z(\phi,\psi) + k_x E_{p,x} f_x(\phi,\psi) + k_y E_n f_y(\phi,\psi) \quad (4)$$

The sum of these two voltages is the total response of the system to the existing field.

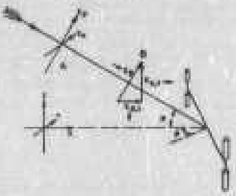


FIGURE 12. Resolution of electric field at collector into components parallel with and perpendicular to plane of incidence.

The factors h and k for the various components are to be obtained empirically, while the directivity functions are deduced theoretically from a knowledge of the configuration of the system. One or two of the h factors in the antenna response may be zero or negligible; for vertical dipoles, for example, h_x and h_y are zero, simplifying the situation. For the undesired response all of the k's may be present; often two are sufficient to describe conditions.

With respect to the directivity functions, the NBS procedure is to determine the dependence on ϕ by measurements at horizontal incidence, while the relation to ψ is determined from a knowledge of the configuration of the systems. These directivity functions are quite general and may be expressed in complex form to account for the phase of each term. They are sufficiently general to permit inclusion of the effect of the field set up by reflection from the ground. The equations, therefore, when expanded to include these factors, describe completely the response of a system under any condition of wave arrival, or ground conditions. This response will depart from the ideal desired response because of the undesired pickup present; an analytical comparison of the actual with the ideal response enables the determination of the polarization errors of the system. Generally, the phase modifications undergone by the various induced voltages through the mechanisms whereby they are induced and transferred from the responding elements to the utilization circuits are not known, nor are they readily determinable, and as a result, the complete equations may not be written explicitly to include them. Nevertheless, a knowledge of the law which the system follows enables the assignment of values to these unknown phase angles such as to make the polarization error a maximum, and thus set an upper limit to the polarization error possible for a particular condition of the downcoming wave. A plot of these maxima over a representative range of elevation angles at an appropriate ratio, say one-to-one, of the parallel and perpendicular field components gives a complete picture of the performance in this range, and may be used for comparison with other systems of the same or different type.

Application to Adcock Antenna

The first step in applying this method, namely, the determination of the directive patterns for the three field components, must now be subjected to further examination. For the purpose at hand, this may best be accomplished in conjunction with an illustrative example. A differential Adcock pair of the elevated H type will be considered, since the NBS report referred to treats a number of this general type.

The H Adcock consists of two vertical dipoles differentially connected by horizontal feeders. The response of this system can be conveniently considered as resulting from a combination of the desired response of the two vertical dipoles and the undesired response of the horizontal feeders. The directivity function for the two dipoles is known accurately on theoretical grounds for fields of any polarization. Obviously $E_{p,z}$ is the only component capable of inducing a voltage in either dipole, since $E_{p,x}$ and E_n are always directed at right

angles to the length of the dipoles. Therefore h_x and h_y are each zero, and terms containing them are eliminated from the response equation.

The horizontal feeders may be replaced, as a first approximation, by a short dipole along the line of the feeders. The directivity function for this dipole is known from theoretical considerations. $E_{p,x}$ is always normal to the direction of this dipole, and can induce no voltage in it; k_x is therefore zero. Further, the maximum response to a unit field of E_n (at $\phi = 0°$), must equal the maximum response to a unit field of $E_{p,z}$ (at $\phi = 90°$, $\psi = 90°$), since in each case the direction of the electric field is parallel to the dipole in question. Therefore the response coefficients k_x and k_y are equal, and the measurement of one is sufficient to establish the other. Evidently both h_x and k_y may be separately determined by measurements at horizontal incidence, and their ratio obtained. Since the complete directive pattern is known, it need not be measured; the NBS procedure is to measure the patterns due to E_n and $E_{p,z}$ at ground level, probably as a verification of the assumptions. The method is thus seen to be substantially an indirect one, in that the polarization error is not measured directly, but is deduced from theoretical considerations and observed data.

It is interesting to consider the situation resulting if in the preceding example it were not possible to assign on theoretical grounds a directivity characteristic to the element responding to the undesired field components E_n and $E_{p,z}$, that is, if knowledge of the behavior of the feeders is insufficient to permit the valid substitution of a simple dipole. It would then become necessary to establish this directivity by empirical methods. A series of measurements would be made, starting for example with E_n fields. The response through 360° in azimuth could be measured for an angle of elevation equal to zero. In carrying these measurements to elevated angles, however, a fundamental difficulty would arise due to reflections from the ground. When a ground-reflected wave exists (and this may even apply in an elevated system to the measurements at zero angle of elevation), there are in effect two waves present, differing in magnitude, phase, and direction of arrival. Moreover, the response of the system to waves from the two directions may introduce additional phase and magnitude changes. Obviously a single figure, the resultant output voltage, is not sufficient to determine uniquely the response in the desired direction. Nor would it be valid to take the downcoming wave, compute the magnitude and phase of the reflected wave, and add the two vectorially in time and space at a point of the system, unless it can be assumed that the point adequately represents the system for the two waves in question, i.e., that waves do in effect act on the system at the point, and nowhere else. For example, if, instead of a single horizontal dipole representing the feeders, it were necessary to substitute two parallel dipoles lying in the same horizontal plane the addition of the direct and reflected waves at one would, in general, not hold for the other, since the path differences in the two cases are not the same for a nearby signal source. This assumption concerning the configuration could not be made, as the configuration itself is to be determined by the measurements. Should an attempt be made to carry the investigation on for the other two field components, difficulties of the same nature would exist and, in addition, other complications would be found. The $E_{p,r}$ and $E_{p,z}$ components are not separable; the plane of incidence and direction of propagation determine uniquely the direction of the E_p vector. Its resolution is useful for analysis, but cannot be accomplished physically so as to eliminate one or the other of the components. As an alternative, the method might be modified to measure the total response to the E_p field, rather than the response to its two components. The effect of the components could be deduced, if the phases of the resultant voltages were known. These, however, cannot as a rule be determined. Moreover, in a reasonably good collector, the desired response of the dipoles to $E_{p,x}$ would almost completely obscure the response to $E_{p,z}$, unless a very high precision were attained in the techniques of the measurement. The desired response could not well be eliminated, since the dipoles, while not responding to the undesired components directly, may be, and usually are, an element in the transfer system from a responding member to the util-

ization circuits, because of radiation or reactive coupling. For example, a member responsive to $E_{p,e}$ may reradiate a component parallel to the dipoles and thus induce a voltage; removal of the dipoles, or short circuiting them would remove this undesired effect; the total effect of $E_{p,e}$ could not then be determined.

Some consideration might be given to a method of evaluating the errors on the basis of the primary components alone, without attempting the more or less artificial resolution of the E_p field. While this might conceivably be possible, the other difficulties mentioned would still be present; in addition to the complicating presence of the ground-reflected wave, the phase angles of the voltages induced by the two components of the parallel field would remain unknown. In itself, this would be of no consequence, since the effect of the whole E_p field is being investigated. However, while the two components are always in phase in a downcoming wave, this may not be the case when the combined direct and reflected waves appear at the collector. As a result, the analytic separation would still appear to be necessary to predict the behavior for any ground conditions.

Free-Space Pattern for Systems Using Reflectors

By positioning the reflector near, and parallel to, the ground, the latter becomes in effect an extension of the reflector; undesired ground reflections would be eliminated. The signal source could be placed directly above the reflector at a suitable height, and either moved through arcs corresponding to azimuth and elevation, or left stationary, and the reflector rotated as required. The objections to this method are that the array would have to be dismantled from its normal operating position, and special gear constructed to obtain the required rotation of the screen, or motion of the source. Further, a legitimate doubt would always exist concerning the equivalence of the undesired response in the operating and measuring positions, since the feeders could not be identically disposed in the two cases. The uncertainty caused by the change in phase of the $E_{p,s}$ and $E_{p,e}$ fields in the presence of the ground would still remain.

These possible procedures have been touched upon not so much to examine their merits, but rather to bring out the difficulties of the method when recourse must be had to an experimental determination of the three dimensional response of an array to certain field components, when the directivities involved cannot be assumed a priori from the configuration of the array. Even in the case of so simple a collector as an H Adcock, it is open to some question whether a successful experimental determination can be made. On arrays of the type studied in this project, the configuration of the elements which may respond to undesired field components is considerably more complex, and the difficulties increase accordingly. The desired response of the dipoles cannot be predicted to a high degree of accuracy because of the presence of an imperfect reflector; to this time a reasonably accurate expression for the current distribution in a cylindrical dipole of large transverse dimensions has not been obtained. Any assumption as to the total response of the system was out of the question.

If consideration be given to the NBS method, its elegance and utility are seen to reside in its ability to assess in terms of simple and readily determinable parameters, a phenomenon which is at best very complex. One of its outstanding advantages in practice is that the measurements are so made as to avoid the complicating influence of the ground, that is, at horizontal incidence. In any of the procedures mentioned above, this advantage would be lost; the measurements would be time consuming and difficult, if at all possible, requiring, in the case of arrays studied in this project, fields of high purity of polarization; the final results would be indirect and subject to question on this ground.

As mentioned previously, an attempt was made to apply the NBS method during the course of the project. Some of the earlier results of measurements on the response of the V-I array to horizontally polarized waves indicated that no simple space pattern could be presumed. The lack of a reliable field-intensity meter hampered the work considerably. Resort had to be made to the array under test for field-intensity comparisons. This was done by orienting the dipole either vertically or horizontally

as required, and assuming the same effective height for the two conditions. Rejection ratios were specified in the bearing direction and are given below. These ratios were found to be of considerable value as an indication of the progress of the work, since any substantial improvement was usually accompanied by smaller measured errors.

Poast Method of Measuring Polarization Errors

The final method adopted for the measurement of polarization errors was one originally intended to verify results of the indirect method. It was originated by L. M. Poast of the National Bureau of Standards, and consists of a means of producing a field polarized so as to have equal components in, and perpendicular to the plane of incidence, with a continuously variable phase adjustment between these components. This is accomplished by exciting three mutually perpendicular electric dipoles from the same shielded source. One dipole, used as the axis of rotation, is horizontal and at right angles to the direction of propagation of interest. The plane through the remaining two dipoles is vertical, and coincides with the plane of incidence at the receiving antenna. The horizontal dipole and one of the dipoles in the vertical plane are fed in phase, and the remaining vertical dipole feed is displaced 90° in time phase by an artificial quarter-wave line. The two dipoles in the vertical plane produce a uniform field in that plane. The phase varies uniformly with angular position in that plane, and the magnitude of the resultant field is equal to the maximum field due to either dipole. There results, therefore, a field as specified above, with a parallel component E_p whose phase may be varied uniformly by rotating the system about the horizontal dipole as an axis. Figure 13 shows the unit which was designed to operate over the 150- to 300-mc band. The dipole extensions are interchangeable with units of other lengths and telescope into the oscillator housing for accurate adjustment to frequency.

When this system is used for the measurement of polarization errors, it is supported so as to permit rotation about the horizontal dipole, then elevated to the desired height. The errors are observed as the assembly is rotated through 360° by means of control cords. The maximum error is noted, as well as errors at uniform angular intervals. The maximum error then is the maximum possible for a one-to-one downcoming field at that elevation angle, receiving antenna elevation, and ground constants. The measurement of the polarization errors is direct, and is, therefore, not open to those objections which are based on the indirectness of a method. Three factors nevertheless may be questioned. The first is the validity of the results based on radiation from a nearby transmitting system. At the lowest frequency in question, the distance between the receiving and transmitting points is about 15λ along the ground and 18λ at the maximum elevation of 34°. It is believed that this is adequate to produce a substantially plane wave front at the receiver. The effect of the surface wave is greatest at low angles of elevation; it may be neglected at the higher angles where the polarization errors reach their maximum

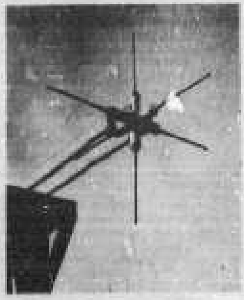


Figure 13. Variable phase polarization transmitter with removable dipole extensions which telescope into oscillator housing for accurate adjustment to frequency.

values. The second objection is that measurements are made with the receiving antenna array at a fixed height above ground, and may not represent the worst point of operation at all frequencies. It was not feasible to construct elevating gear for this work. To overcome this objection to some extent, each curve of polarization error reproduced in this report has a section showing the ratio of E_n to the $E_{p,z}$ components at the receiving antenna through the range of elevation used. From this it may be determined whether a specific error was obtained under favorable or unfavorable ground reflection conditions, and the extent of discrimination against one or the other field component. The third objection is that the information obtained covers only a limited range of conditions, and does not specify the complete performance.

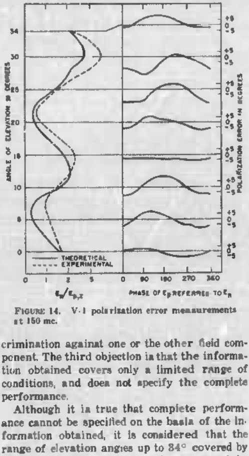


FIGURE 14. V-1 polarization error measurements at 150 mc.

Although it is true that complete performance cannot be specified on the basis of the information obtained, it is considered that the range of elevation angles up to 34° covered by the data is wide enough to include most of the practical conditions of operation likely to be encountered. In the upper end of the v-h-f and the u-h-f[b] bands, high-angle waves originate generally from elevated sources—aircraft transmitters primarily. In homing operations of friendly aircraft, for example, angles of elevation over 34° will rarely be found beyond a horizontal distance of the order of one mile.

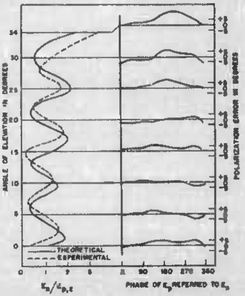


FIGURE 15. V-1 array, polarization error measurements at 300 mc.

The use of this rotating phase arrangement requires eight or more observations at each angle of elevation. To expedite measurements for day-to-day comparison of results, the simple elevated dipole, tilted ±45° from the vertical, was resorted to. This method, while not giving the maximum error, yielded results of sufficient significance to be quite adequate for the purpose, and the measurements were readily repeatable after several days' lapse. It is interesting to note that the rotating phase method rather consistently gives a maximum error

[b] By rather common agreement the various bands are regarded as including the following frequencies: v-l-f, 3-30 kc; l-f, 30-300 kc; m-f, 300-3,000 kc; h-f, 3-30 mc; v-h-f, 30-300 mc; u-h-f, 300-3,000 mc; s-h-f, 3,000-30,000 mc.

through the range of elevations that is about 50 per cent in excess of the errors measured with the tilted dipole. The latter results are given in Tables 1 and 2. The errors obtained using the variable phase method at 150 mc are shown in Figure 14, and at 300 mc are shown in Figure 15.

4.3.7 Selection of Optimum Height

The selection of a suitable height for a d-f array should be guided by two performance considerations, aside from the purely mechanical ones involved in the design of a satisfactory elevated rotating mount.

In the upper end of the v-h-f range, the tendency of electromagnetic waves to propagate along optical paths becomes evident, and this tendency becomes more marked as the frequency is increased. In the u-h-f range the paths are essentially optical. The curvature of the earth therefore limits the distance which may be covered, since the optical path is a straight line. The phenomenon of refraction in the atmosphere modifies this condition somewhat, as the path followed curves back toward the earth relative to a straight line tangent to the earth. Quantitatively the effect may be accounted for by assuming a radius for the earth in excess of its actual radius. On this basis and the geometry involved, the distance to the effective horizon is given in terms of the height by

$$d_{\text{miles}} = \sqrt{2h_{\text{feet}}}$$

wherein the effective radius of the earth is taken as 1.32 times the physical radius. The obvious conclusion to be drawn is that to obtain maximum range, as great a height as practicable should be used for the direction-finder array.

The second consideration influencing the choice of height is the effect on polarization errors. Due to interference phenomena between the direct and ground-reflected waves, a standing wave pattern of field intensities is set up along the vertical line over a point. This pattern is different for the perpendicular and parallel field components, so that the ratio of the two varies with elevation over the point in question; therefore relatively large suppression of one or the other component is possible. The degree of suppression is dependent on the elevation, the electrical characteristics of the ground, and the angle of elevation of the downcoming wave.

The interference pattern is a result of the phase difference existing at a point between the direct and reflected waves. This difference is made up in part by the phase shift occurring at reflection; the remainder is due to the difference in the paths traveled by the two waves. The corresponding phase difference for the latter is given by

$$\Delta = \frac{4\pi h \sin \psi}{\lambda}$$

where Δ is the phase difference in radians, h the elevation of the point in question, ψ the angle of elevation of the arriving wave, and λ the wavelength. The difference is seen to increase directly as the height and the sine of the angle of elevation. The change of phase with ψ is consequently more rapid as h is increased.

The phase change occurring at reflection is given by the appropriate Fresnel plane wave reflection coefficient for the parallel and perpendicular cases. These are

$$R_p = \frac{\epsilon \sin \psi - \sqrt{\epsilon - 1 + \sin^2 \psi}}{\epsilon \sin \psi + \sqrt{\epsilon - 1 + \sin^2 \psi}}$$

$$R_n = \frac{\sin \psi - \sqrt{\epsilon - 1 + \sin^2 \psi}}{\sin \psi + \sqrt{\epsilon - 1 + \sin^2 \psi}}.$$

Here the complex dielectric constant

$$\epsilon = \epsilon_0 - j\epsilon_1$$

ϵ_0 = dielectric constant in esu;
ϵ_1 = $2c\lambda\sigma$;
c = velocity of light in cm per sec;
λ = wavelength in cm;
σ = conductivity in esu;
R_p = parallel reflection coefficient;
R_n = perpendicular reflection coefficient.

Both the magnitude and phase angle of the coefficients for the two cases vary in a different manner with the angle of elevation. The phase angle of R_n remains nearly constant, while the phase of R_p undergoes approximately a 180° change as the angle of elevation changes from 0° to 90°. The overall effect is that the ratio of the E_n resultant to the E_p resultant varies through wide limits along a vertical line over

a given point. A plot of this ratio is given in Figure 16, against height in terms of wavelength, for three values of the parameter ψ. These curves are computed for a complex dielectric constant of $10 - j1$, and correspond to the ground constants at the Medford site for

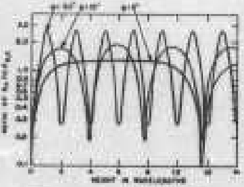


FIGURE 16. Ratio of E_H to E_V versus height.

the low-frequency end of the range (150 mc). The imaginary component is small enough to be neglected, and decreases with increasing frequency. The dielectric constant of 10 may be taken to represent average ground conditions in the frequency range investigated. An examination of these curves leads to two conclusions: first, the ratio of the horizontal to the vertical field intensities is consistently small only at elevations less than $\lambda/4$. At 600 mc this represents a height of about 5 inches, and at 150 mc, 20 inches, values too small to be usable. For elevations in the usable range, say over 6 feet, the height would represent several λ at the higher frequencies. Second, consistently small ratios for different elevation angles are not possible for a given height over $\lambda/4$ even at one frequency. For example, all three curves go through a minimum in the vicinity of 6λ; at intermediate or other angles, this would not necessarily be the case. Reference to the polarization errors of the V and flat arrays shows that at the low-frequency end the errors are the greatest, and these occur at high elevation angles. If one assumes that the maximum angles encountered in practice are in the vicinity of 30° to 35°, it would be possible to select a height giving favorable ratios near the low-frequency end of the band for high angles, but the favorable ratios would not hold elsewhere. In the absence of elevating gear and means for determining the angle of elevation of an arriving wave, it would appear that a selection of height based on maximum range, and completely random as far as the present consideration is concerned, is as likely to result in satisfactory operation as would a height selected for a particular set of conditions.

The NBS report[3] has data similar to Figure 16. The latter is somewhat more general in that elevations are given in terms of wavelength, and the ground conditions specified in terms of the complex dielectric constant. Figure 16 is therefore usable directly at any frequency for a complex dielectric constant of $10 - j1$.

4.2.3 Array Gain

To measure the gain of the V-1 array, a configuration was required which would eliminate ground reflection effects. A simple manner of achieving this consists of performing the measurement in the vertical direction. The reflector in question is placed parallel to the ground, with the dipole above the reflector. A device to indicate relative field intensities is placed directly above, and elevated to a height great enough to eliminate spurious proximity effects. The antenna in question is excited with a power oscillator, and the field intensity so obtained is compared to that produced by a resonant half-wave dipole, $\lambda/4$ above the screen, in exactly the same position. Relative input powers are measured by standing-wave equipment for the two cases. The relative gains of the two arrays are then computed.

The absolute gain of the standard antenna may be obtained theoretically and, together with the relative gain, enables the determination of the absolute gain of the array being measured. Figure 17 shows the gain characteristics of the V-1 array over the entire frequency range, the standard of comparison being a hypothetical isotropic or nondirectional antenna. For comparison with a half-wave dipole in free space, the values given in this curve should be reduced by 2.14 db. This

figure represents the gain of a half-wave dipole in free space over an isotropic antenna. The variation of gain with frequency is seen to be slow for this array.

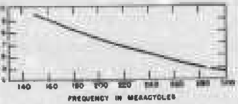


FIGURE 17. Gain of V-1 array compared to hypothetical isotropic or nondirectional antenna.

V-1 Array Used as Direction Finder

Following the decision to use a balanced dipole system, an array was set up with balanced two-wire lines connecting each antenna to the switch, and a twin-conductor lead from the switch to the receiver. The performance of this system was fair, but it was obvious that there was considerable signal pickup due to the dipoles and feeders responding as a unit to fields between them and ground, and that it would be necessary to install the equivalent of a balanced and electrostatically shielded transformer. A suitable design was selected, utilizing resonant lines; the principles of operation and design formulas are given below. One transformer was placed behind each screen at the point where the dipole transmission lines pass through the screens. Rejection ratios (corrected for curvature of receiver input/output characteristics) measured prior to the installation of these transformers were approximately 10/1 at 150 mc, and 5/1 at 300 mc. The use of the transformers improved the ratios to about 40/1 through the frequency range. The measured maximum polarization errors before installation were approximately 50°, and these were reduced by a factor of two through the use of the transformers. The remaining errors were considered too high, and further studies were undertaken in an attempt to obtain a reduction.

The errors mentioned were noted when the system operated as a switched-lobe direction finder, that is, one in which amplitude comparison is made by successive observations on each screen. In order to check the electrical balance between the screens simultaneously, the two arrays were differentially connected, in which case phase and amplitude balance are indicated by a null. The errors noted with this arrangement were lower by a factor of perhaps three-to-one, indicating good electrical balance. The reason for this wide discrepancy is not completely understood at this time, but is mostly likely due to the inherently balanced nature of the differential system as compared to the dissymmetry existing when only the left or right half of the array is observed at one time, which condition holds in lobe switching.

Further investigations tended to confirm this explanation. Measurements were made to compare the response of the two halves when the polarization of a horizontally incident wave was varied. With vertical polarization, the response patterns of the two antennas were nearly identical; slight differences were attributed to the outputs resulting from the $E_{y,z}$ component of the ground-reflected wave, adding at different phase angles to the respective $E_{y,z}$ voltages in the two antennas. When the plane of polarization was rotated clockwise as viewed from the receiver, the response of one increased, and the other decreased; counterclockwise polarization produced the opposite effect. This effect was found to depend on the angle between the line of propagation and the normal to the screen. When the screens faced the source, the effect was a minimum.

EFFECT OF SUPPORT POLE

An element of dissymmetry appeared to be the support pole, and its effect was next investigated. The transmitter was polarized horizontally, and its output increased sufficiently to produce an output in the receiver. With this condition, standing waves were noted along all the edges of the screen, except in the vicinity of the support pole. The edges of both screens were then insulated from the pole to obtain a more symmetrical potential distribution. A decided improvement resulted; the response patterns were more nearly alike, and the polarization errors reduced. The same effect was observed with the differential connection before insulating the screens, but while the two

lobes of the pattern using this connection changed in relative size, the position of the null remained substantially unchanged. With the screen insulated, the resulting pattern was symmetrical regardless of the polarization of the transmitter.

Similar observations were made on the V-2 array, but the effect of insulating the screens was much less pronounced; first, because in the bearing position the normals to the screen lie more nearly along the plane of propagation; and second, because the higher gain of the array provides better discrimination against reradiation effects due to horizontally polarized components.

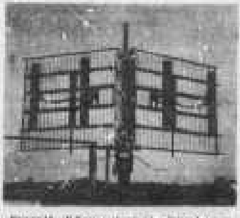


FIGURE 18. V-2 array, showing insulators between screens and shaft.

Figure 4 shows the V-1 array before the screens were insulated from the shaft. The insulating blocks may be seen in Figure 18 which shows the final V-2 array.

4.4 V-2 ARRAY (2 DIPOLES PER REFLECTOR)

The V-2 array is similar to the V-1 in principles of operation, the major point of departure being the use of a broadside array of two dipoles on each reflector. Consequently the general discussions covering the V-1 array are applicable here.

The use of two dipoles as compared with one per screen is advantageous in a number of respects. The gain is increased through improved directivity in azimuth, while the vertical directivity remains unchanged; the differential sensitivity is higher; polarization errors are reduced; the size of the array is substantially unchanged. Figure 18 is a view of the V-2 array with the edges of the screens insulated from the shaft.

4.4.1 Experimental Work

The first array studied had a spacing between the line of dipoles and screen equal to 28.5 cm, the same as was used for the V-1 array. This spacing was maintained through the tests and was considered to be an optimum from the standpoint of gain and impedance characteristics, although more latitude is available in this array than in the V-1 array. The distance between the two dipoles of a screen was made 66 cm, or approximately one-half wave near the middle (225 mc) of the frequency range. A set of operational data was obtained including polar patterns, polarization errors, and gain. The data indicated that this array was considerably superior to the V-1 array primarily because of the improved directivity in azimuth. To increase the directivity further, the spacing between dipoles was increased to 86 cm, representing a half wavelength at 175 mc, without changing the screen dimensions. Observations were made with this spacing, the maximum that the screen will accommodate and still have the required one-eighth wave projection beyond the dipoles. Further increase is not usable, since the spurious side lobes at the high-frequency end become troublesome. Comparative data on the V-1, V-2 (66 cm), and V-2 (86 cm) arrays are considered in Tables 1 and 2. As in the V-1 array, both switched-lobe and differential operation were investigated.

4.4.2 Relative Response in Azimuth

Polar diagrams showing the relative azimuthal response for one-half of the V-2 array with 86-cm spacing are given in Figures 19 and 20 for 150 and 300 mc, respectively, at 0° elevation. The increased directivity over the V-1 array is clearly evident in these diagrams.

TABLE 1. Comparative polarization errors, V-1 and V-2 arrays.

Angle of elevation in degrees	Array	Switched-lobe connection (Error in degrees)						Differential connection (Error in degrees)			
		V-1		V-2 (66 cm)		V-2 (86 cm)		V-1		V-2 (66 cm)	
	Polarization	−45°	+45°	−45°	+45°	−45°	+45°	−45°	+45°	−45°	+45°
0		−3	−3	+10	−3	+2	−3	+0.75	+1.5	0	+1
5		−3	−4	+1	0	+1	−2.5	+0.5	+3.0	−0.5	+2
10		−6	−1	−3	+5	−4	+2	+0.5	+3.0	+1	+1
15	150 mc	−5	+2	+12	−4	+4	−5	+0.5	+0.5	0	0
20		−11	+1	+2	+1	+1	−0.5	+0.5	+2.5	−1	+1
25		−15	+3	−7	+7	−4	+4	+1.0	+4.0	0	+2
30		−16	+13	+12	+4	+2	0	+4.5	+2.0	+3	0
34		−10	+7	+12	0	+5	−3	+1.5	+2.0	−0.5	0
0		−0.5	+1	−0.5	0	+0.25	+0.5	0	+0.5	+0.25	+0.25
5		−3	+2	−0.25	−0.25	−0.75	0	+0.2	+0.2	0	0
10		−1	+2	−1.0	+0.5	−0.5	+1	0	−0.2	0	+0.5
15	300 mc	−0.5	+3	−0.5	+0.25	−0.75	+1	0	0	0	+0.5
20		+0.5	+1	0	0	−1	+0.75	+0.2	−0.2	0	0
25		+1	+2.5	0	−0.25	−0.5	+1.25	+0.5	−0.2	−0.5	0
30		−3	+4	−4	+0.5	−1.5	+2	+0.6	+0.5	−1.0	+0.25
34		−3	+0	−2	+1.0	−1.5	+2	−2.5	−3.5	0	0

TABLE 2. Summary of maximum errors.

Frequency	Switched-lobe connection (Error in degrees)			Differential connection (Error in degrees)	
	V-1	V-2 (66 cm)	V-2 (86 cm)	V-1	V-2 (66 cm)
150 mc	16	12	5	4.5	3
300 mc	9	4	2	3.5	1

The directivity increases with frequency over the range illustrated; this is in opposition to the behavior of the V-1 antenna, where maximum directivity occurs at the low-frequency end. Response diagrams for 66-cm spacing are not given; these are very closely similar to the

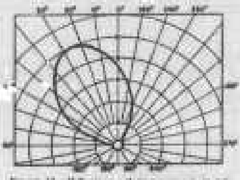


FIGURE 19. V-2 array, relative response in azimuth, 0° elevation, 150 mc.

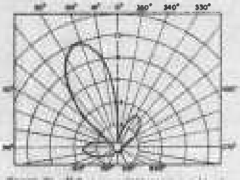


FIGURE 20. V-2 array, relative response in azimuth, 0° elevation, 300 mc.

ones shown, but are slightly broader. The present diagrams were obtained before the screen angles were adjusted for the optimum position. This may be noted in the 300-mc diagram where the response on bearing is too low. The patterns should, therefore, be rotated approximately 6° toward the zero azimuth line to correspond to optimum setting. The relative response in elevation for the lobe-switching connection is given in Figure 11.

4.4.3 Impedance Characteristics

Figure 21 is a plot of the impedance characteristics of the V-2 array at the balanced-unbalanced transformer, and includes the effect of the latter, as well as of the transmission

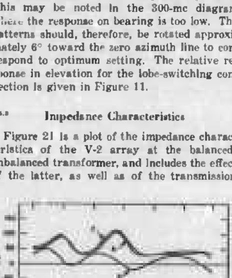


FIGURE 21. V-2 array impedance characteristics. Dipoles spaced 86 cm (1/2 at 175 mc); spacing to reflector is 29.5 cm (1/4 at 250 mc).

lines between it and the dipoles. The impedance is comparatively uniform through the 140- to 300-mc frequency range, with the reactive remaining less than the resistive component through the range. The geometric mean of the minimum and maximum points is approximately 57 ohms; therefore, standard 60-ohm cable may be used without additional matching transformers. The impedance mismatch when using 60-ohm cable does not exceed two-to-one, and is considerably less through most of the range.

4.4.4 Polarization Errors

Comparative data on polarization errors of the V-1 and V-2 arrays are given in Table 1 for both the switched-lobe and differential connections. (The latter connection is more fully discussed later in this chapter.) These were obtained with a tilted dipole transmitter. The tilt in this test is always about a horizontal line lying in the plane of incidence, i.e., the dipole always lies in a vertical plane normal to the plane of incidence, and, therefore, the E_v component of the downcoming wave is less than the E_h component at elevated angles. The ratio of the two is nearly proportional to the cosine of the angle of elevation; i.e., unity at horizontal incidence, and dropping to 0.83 at 34° elevation. It is to be noted that the differential connection is better by a factor of three or four to one compared to the corresponding switched-lobe array in regard to polarization errors. Nevertheless, the errors of the V-2 array using 86-cm spacing between the dipoles are quite low for lobe-switching operation. For rapid comparison, Table 2 lists the maximum errors found in Table 1.

It is evident from these tables that there is a progressive improvement in polarization error performance as the azimuthal directivity is increased. Consequently it is reasonably safe to predict that still greater improvement is possible if arrays of greater directivity are used. Because of size, their use would probably be limited to permanent or semipermanent installations. It is pertinent to observe that as a result of increased directivity, searching becomes more difficult, since high response is limited to a narrower azimuthal sector. Excessively directive arrays may require the use of subsidiary searching equipment.

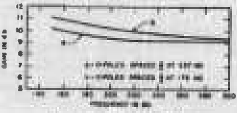


FIGURE 22. Gain of V-2 array.

4.4.5 Gain of V-2 Array

Measurements of gain on this array were made at both the 66-cm and 86-cm spacing, and are given graphically in Figure 22. The im-

provement obtained by the wider spacing is greatest at the low-frequency end; at the high-frequency end the appearance of side lobes limits the possible improvement. As in the case of the V-1 array, the standard of comparison in the curves is a nondirectional or isotropic antenna. For comparison with a half-wave dipole in free space, the gain taken from these curves should be decreased by 2.14 db.

4.8 FLAT ARRAY

4.8.1 Theory of Operation

The principle of operation of the flat array in which the directivity pattern is shifted in azimuth by a change in phase of some of the elements may be readily seen from the following considerations: If in Figure 23 we have two electric doublets, 1 and 2, in a radiation

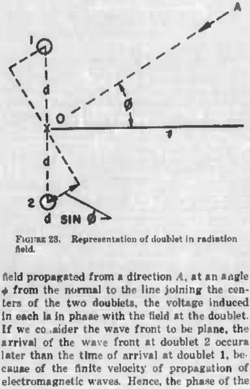


FIGURE 23. Representation of doublet in radiation field.

field propagated from a direction A, at an angle ϕ from the normal to the line joining the centers of the two doublets, the voltage induced in each is in phase with the field at the doublet. If we consider the wave front to be plane, the arrival of the wave front at doublet 2 occurs later than the time of arrival at doublet 1, because of the finite velocity of propagation of electromagnetic waves. Hence, the phase of the induced voltage in doublet 2 lags the voltage in doublet 1 by βx, where x is the additional distance traveled, and β is the phase constant of free space, equal to $2\pi/\lambda$, λ being the free space wavelength. It is convenient to refer phases to the field at 0, the center of the line joining the two doublets. Then if d is the distance from this center to either doublet, the voltage induced in 1 leads the field at 0 by an angle $2\pi d \sin \phi/\lambda$, while that in 2 lags by the same angle. A vector diagram is shown in Figure 24, where

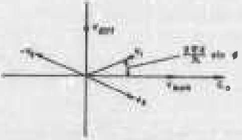


FIGURE 24. Vector diagram of doublet voltages.

E_0 is the field at 0, V_1 is the voltage induced in doublet 1, and V_2 the voltage in doublet 2.

Either the vector sum or difference of these two voltages, which are also shown in the diagram as V_{sum} and V_{diff}, may be utilized. The salient fact revealed in this diagram is that the sum voltage is in phase with the field at 0, while the difference voltage is displaced 90° in time-phase from both E_0 and the sum voltage, for equal magnitude component voltages V_1 and V_2.

$$V_1 = kE_0\left[\cos\left(\frac{2\pi d}{\lambda}\sin\phi\right) + j\sin\left(\frac{2\pi d}{\lambda}\sin\phi\right)\right]$$

$$V_2 = kE_0\left[\cos\left(\frac{2\pi d}{\lambda}\sin\phi\right) - j\sin\left(\frac{2\pi d}{\lambda}\sin\phi\right)\right]$$

where k is a proportionality factor. Therefore

$$V_{sum} = 2kE_0\cos\left(\frac{2\pi d}{\lambda}\sin\phi\right)$$

$$V_{diff} = j2kE_0\sin\left(\frac{2\pi d}{\lambda}\sin\phi\right).$$

The sum voltage is thus real, and in phase with the field E_0, while the different voltage is imaginary, and therefore displaced 90° from E_0 and V_{sum}.

Typical directional patterns for the differential and additive cases are shown in Figure 25 B and C for the case of spacing between the doublets of the order of a half wavelength.

In the differential case, the resultant voltage is zero when ϕ is zero, while in the additive case the voltage is a maximum. In both cases, there is a reversal in phase where the resultant passes through zero. The axes of maximum response for the two cases are displaced from each other 90° in azimuth. A 90° phase change introduced in the output of either one or the other will bring corresponding lobes in phase, but will not change the space pattern. Therefore the voltages from two pairs, one additive, and the other differential, may be added, and the resultant space pattern will be rotated in azimuth. In Figure 26, A shows two such pairs, disposed along the same line; at B is shown the pattern of the differential pair, with an advance in phase of 90° introduced in its output, while at C the sum pair is shown unchanged. The resultant pattern at D has its line of maximum response along a line intermediate between the lines of the individual maxima. If we consider the axis of the sum pattern as a reference direction, then the resultant pattern has been rotated clockwise. Obviously a reversal in phase of either pair will rotate the resultant counterclockwise by a like amount.

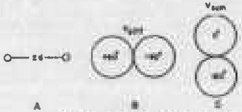


Figure 25. Directional patterns for doublets separated λ/2; B shows differential voltage pattern; C shows sum voltage pattern.

It is interesting to observe a close similarity between the action of this system and that of switched cardioids, obtained, for example, by a loop and sense antenna. In the latter case the voltage induced in the loop is proportional to the time derivative of the magnetic field, and since this latter is in phase with the electric field, the induced voltage is displaced 90° in phase from the electric field. The sense antenna voltage is in phase with the electric field. Or, alternatively, the vertical members of the loop may be considered electric doublets, differentially connected by means of the horizontal members, yielding the same result.

The use of a reflector behind the line of doublets removes one lobe of the response pattern, leaving one point of intersection when the pattern is alternately rotated clockwise and counterclockwise. This intersection represents equal response to the same wave by each array and may be used as a bearing indication.

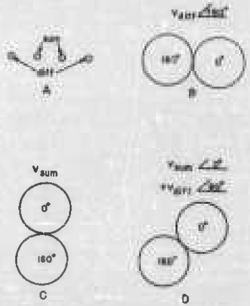


Figure 26. Directional patterns of two sets of doublets in line with voltage of differential pair advanced 90°.

4.5.2 Physical Arrangement

The actual collector system developed follows basically the scheme outlined above. Since a wide band is covered by the antenna system in question, quantities given in terms of wave-

length refer to the wavelength at the arithmetic mean frequency unless otherwise specified. Figure 27 is a schematic diagram of the array. For reasons of symmetry, one pair of dipoles is placed between the dipoles of the other. On the figure, the outer pair, spaced one wavelength, are differentially connected, while the inner pair, spaced $\lambda/2$ apart, are connected additively. All dipoles are placed 28.5 cm, or approximately $\lambda/4$ from the reflector at 260 mc. Balanced feeders run from each dipole to

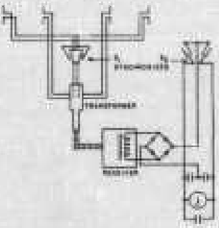


FIGURE 27. Schematic diagram of [illegible] array.

the screen, and through the screen to the switch or transformer, as the case may be. Switching may be accomplished in either pair; in the final experimental model the outer pair were switched. The feeders to the outer pair exceed in electrical length the ones to the inner pair by approximately $\lambda/4$ to introduce the phase lag of 90° required.

4.8.3 Choice of Electrical Elements

To cover a two-to-one frequency range satisfactorily, the electrical characteristics of the various elements making up the array must be carefully chosen. Brief considerations will indicate the large number of parameters, each of which individually affects the performance, and many of which are interdependent. Theoretically, for nondirectional "point source" elements, the array will produce an ideal directional pattern having no spurious response lobes when the phase shift introduced artificially is exactly 90° and the amplitudes utilized from the center pair and outer pair respectively bear a ratio of two to one. Since it was considered undesirable to control the relative amplitudes of the two pairs, investigation indicated that good results could be obtained with a one-to-one amplitude ratio, allowing the ideal 90° phase shift to change with frequency from 60° at the low-frequency end, through 90° at the center frequency, to 120° at the upper end of the frequency band. Limiting the phase shift to this range of values and maintaining a one-to-one amplitude ratio through th frequency range, presupposes resistive dipole elements matched to the transmission lines, with no mismatch at the junction or other points of the system.

Reference to dipole impedance characteristics, Figure 6, obtained during the development of the V-1 antenna system will show that it is impossible to obtain a uniform resistive characteristic over the frequency range. While quite good standing-wave ratios are obtainable using one or more dipoles feeding suitable lines in the V-1 system, where phase shift is of secondary consequence, in the present case, where spurious phase shifts may make the system inoperative, attention must be given to all factors which can contribute to phase and amplitude variations. Since the presence of mutual radiation impedance between a dipole and its image, and between two dipoles tends to increase the variation, over a range, of the total impedance of a dipole, spacings between dipoles and from the reflector must be chosen to keep these mutual impedances at as low a value at the lowest frequency used as is consistent with other requirements. This means that spacings between dipoles and from dipole to screen should be large in terms of wavelength at the lowest frequency used.

Conflicting with this requirement is the phenomenon of spurious response lobes appearing at the high-frequency end of the band when spacings are of the order of one wavelength or more. The choice of these spacings must therefore be a compromise based on these two limiting factors.

For the same reasons, the self-impedance of the dipoles should be as uniform as possible and essentially resistive. The dipole length to diameter ratio (7.5/1) used in the V-1 system, offers a fair approximation to the ideal condition. A better approximation is not possible without increasing the diameter to a size considered excessive for portable use. For fixed-station direction finding, however, modifications along these lines should produce a collector system capable of more uniform performance through a two-to-one frequency band.

Transmission Lines

At the center frequency, the transmission lines to the outer pair exceed the inner lines by nearly one-quarter wave in order to produce the required phase shift. As outlined previously, the ideal phase difference is 90°. A quarter-wave excess in tranmission lines produces this phase difference when no impedance mismatch occurs in the system. When terminated by actual dipoles, however, whose impedance varies through the band and is partly reactive excepting at a few points in the band, this condition does not hold. Further, the quarter-wave excess produces a transformation in impedance in addition to that occurring in the shorter line. Therefore, at the junction point, the impedance presented by one set of lines, terminated by its pair of dipoles, is generally different from that of the other. As a result, both the phase and amplitude of the two currents in the load are modified by an undesired amount. This last factor should, however, be qualified to this extent, that the amplitude modification may be in a direction to approach a two-to-one ratio, which is preferable to the one-to-one ratio, and also, when operating away from the center frequency, where the nominal phase shift is more or less than 90°, the change may tend toward the 90° value desired. Both, of course, may move in the wrong direction. A result which is unqualifiedly desirable is that at the center frequency, the impedance transformation of the lines to one pair of dipoles is the inverse of the transformation in the lines to the other pair, referred to the characteristic impedance of the line. This means that the reactances are of opposite sign and at least partially cancel. Off the center frequency, while the transformations are not exactly inverse, they are nearly so, and reactance cancellation still occurs. The impedance of the system as a whole consequently has small phase angles through most of the range, as shown in Figure 28.

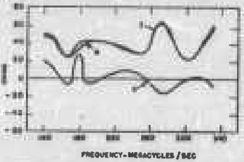


FIGURE 28. Flat array impedance characteristics.

These considerations should make it evident that in addition to the excess line in one branch, the characteristic impedance and total length of lines must be properly chosen. The number of impedances available in standard solid dielectric low-loss high-frequency lines is limited. The lines used have an effective impedance that approximates the geometric mean of the impedance range of each dipole; this minimizes the variation of the transformed impedances. The total length of lines should be kept as low as possible to minimize losses and undesired pickup, with this reservation, however, that they should be so selected as to avoid quarter-wave transformations at points where the dipole impedance departs farthest from the characteristic impedance of the line. This is particularly the case at the low-frequency end of the band. Should a quarter-wave transformation occur here on one of the lines, the other may be near a half-wave transformation point; the latter will remain substantially unchanged, while the former will be raised in impedance by a factor of perhaps three or more. If this happens to be the center pair, its output current will be reduced by a factor of three or more, thus departing by a factor of six from the ideal two-to-one ratio.

The physical disposition of the elements of the array sets a lower limit on the usable

length of transmission lines. This minimum is somewhat more than one meter from the transformer, through the switch, to each outer dipole. Since the velocity of propagation in the 50-ohm polyethylene cable used is approximately 64 per cent of the velocity of light in free space, the actual electrical length is greater than the mechanical length by a factor of approximately 1.6. For this reason, air dielectric lines could possibly be used to advantage. Furthermore, the attenuation factor of air dielectric lines is generally lower, and more latitude is available in the choice of characteristic impedance. Due chiefly to the ease of adjustment to length, the experimental work on this array was completed using only the solid dielectric cable mentioned. The actual lengths

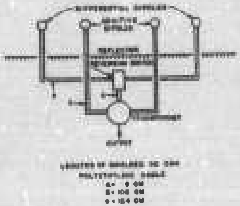


Figure 29. Cable lengths employed in connecting flat array.

used are given in Figure 29. The electrical difference in length is approximately 77°, this figure producing the best performence through the range as determined experimentally. A word of caution is appropriate at this point concerning line lengths. Should it be desiced to duplicate this array, the electrical length of the lines must be accurately set. While not critical the adjustments should be made to within ½ cm or less. The velocity of commercial cable varies between different runs. All commercial cable should, therefore, be measured, and cable preferably from the same run be used on one array.

Balanced-to-Unbalanced Transformer

The transformer for converting the balanced system to unbalanced feed is similar to those used in the V-1 array. A single transformer is quite satisfactory. While reactance cancellation by means of a half-wave series line is possible as in the case of the V-1 array, it may be omitted here with very little change in overall performance. The reason for this is that in effect four dipoles are paralleled (after impedance transformation by their individual lines) at the transformers, resulting generally in a lower effective impedance than the individual dipoles have; the characteristic impedance of the quarter-wave transformer lines is high in comparison with this, and is increased by a factor equal to the tangent of the phase length, so that the effective shunt reactance is high, resulting in a low equivalent residual series reactance.

Reflector Dimensions

The dimensions of the screen used for this array are 120 cm high and 188 cm wide. The spacing between adjacent vertical elements is the same as used in the V-1 array, namely, about $\lambda/20$ at the highest frequency covered. By substituting fine mesh high-conductivity screen, this spacing was found to be adequate in that array. The overall size is about the minimum that can be satisfactorily used. Some improvement in gain at the low-frequency end is possible by increasing the reflector size. For sizes smaller than used, the pattern broadens considerably, resulting in lowered gain.

Relative Response in Azimuth

The performance of this system compaces favorably with that of the corner type using an array of two dipoles per screen. Response patterns for the flat array, at 140 and 300 mc, are given in Figures 30 and 31. While the patterns exhibit considerable variation through the band, as compared to the corner type, the intersection points of overlapping lobes are satisfactory. The adjustment of the system in this respect is very much more restricted than in the V type, where a mere change in the screen angle changes the intersection point. While at any one frequency the pattern may be changed over wide limits by adjustment of

tha line lengtha, this process also changes the pattern through the rest of the range in a different manner. The intersection points should

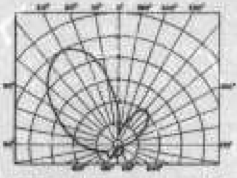


Figure 30. Flat array, relative response in azimuth at 140 mc.

be conaldered fixed, unless variable controls are incorporated in the aystem, and this was ruled out in setting the preliminary scope of work.

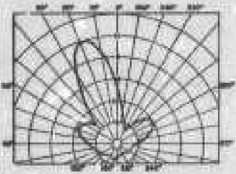


Figure 31. Flat array, relative response in azimuth at 300 mc.

Polarization Errors

The polarization errors of the flat array are quite low. The method of presenting polarization error data is the same as used in the V-1 array. On each graph showing the error under various conditions is placed a curve showing the horizontal-to-vertical field ratios through the range of elevations used. The maximum error observed at 150 mc is 6.5°; at 300 mc the maximum error is 3.5°.

Impedance Characteristics

As indicated above, the approximately inverse transformations occurring in tha transmission lines to the outer and inner dipoles, produce a comparatively high degree of reactance cancellation. The resulting impedance of the system, as seen at the transformer, and including the effect of the latter, is quite uniform, and has small phase angles through most of the range. Reference may be made to Figure 28, which gives the impedance and the resistive and reactive components through the frequency range.

Gain of the Flat Array

As may be inferred from an examination of the relative response patterns, the gain of the flat array is less uniform through the band than the gain of either the V-1 or V-2 arrays. The variation is cyclic, but its magnitude is not large enough to be serious. Figure 32 gives

Figure 32. Flat-array gain characteristic compared to that for nondirective antenna.

the gain over the range, compared, as in the V arraya, to a nondirective, or isotropic antenna. For comparison with a half-wave dipole in free space, figures obtained from this curve should be reduced by 2.14 db.

4.4 SWITCHING AND INDICATING DEVICES

4.4.1 Switches

To switch between antennas of the V arrays and to obtain the required phase reversal in the flat array, a motor-driven switch was developed that has electrical characteristics similar to the transmission lines so that discontinuity of impedance and the resulting reflections are minimized. The desired characteristics are obtained by adjustment of capacitance per unit length to the required value.

The type of switch employed is shown in Figure 33 and consists of two moving contact members and four fixed contacts. The moving contacts are driven through an eccentric ball-bearing race. The fixed contacts are adjustable so that adjustments may be made which permit

FIGURE 33. [illegible] in flat array.

closing the r-f section before closing the meter contacts. Adjustments are also possible which permit opening the r-f circuit before the indicator circuit. This arrangement was found necessary to eliminate transients in the meter because of the antenna make-and-break. The bearings of each moving contact are clamped in rubber pads between bakelite blocks in order to minimize chatter.

Previous experience in the construction of a switch for similar functions showed that the selection of the correct contact material was important. Silver, gold, iron, and several other metals and alloys proved unsatisfactory where extremely low r-f currents were to be broken, even though fair contact pressure was available and the contacts were mechanically wiping. The most satisfactory material found, and one which operates for long periods without trouble from varying resistance, is rhodium.

As used on the V array the switch consists of two sections. The first section switches the leads from each side of the array to the receiver line and simultaneously disconnects the unused half of the array and grounds it by means of back contacts. The second section of the switch consists of a mechanically similar unit connected to operate as a single-pole single-throw switch to couple the receiver output to the indicator bridge.

When used with the flat array, the first section of the switch was modified to become a double-pole double-throw unit with the back contacts insulated from ground and utilized as shown schematically in Figure 27.

The motor used to operate the switch has a 12-volt universal winding, coupled to the switches through a ten-to-one reduction gear. Speed is controlled by a Variac in series with the primary of the supply transformer. Normally the switch is operated at a speed between five and ten cycles per second. Limitations in the maximum speed are purely mechanical. The indicator damping and desired responsiveness set the lower limit to the usable speed.

The shafts of the switch sections are linked together through Oldham couplings. These allow removal of individual switch sections for repair or adjustment and permit proper replacement of the switch without the necessity of resynchronizing, or the reorientation of the shafts, since the latter can be reassembled only in the desired position. In addition, this type of coupling takes up misalignment of shafts.

The motor leads are unshielded and run through the hollow aluminum antenna drive shaft without r-f filtering. The r-f interference caused by sparking motor brushes appeared to be entirely absent at the frequencies used, and no trouble was encountered from this source.

4.4.3 Indicators

The indicator used in the majority of tests consists of a simple zero-centered, 100-microampere d-c meter having rather high electrical damping. The scale is marked off with the letters L-O-R, indicating the direction in which

the array should be rotated in order to obtain the bearing.

The connections to the indicator from the receiver and switch are shown schematically in Figure 34. The use of the capacitors C_1 and C_2 in place of a resistor network is advisable as it allows considerable latitude in the adjustment of the switch contacts. The dwell periods do not need to be equal with this arrangement since it operates similarly to a peak voltmeter.

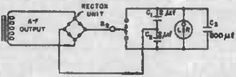


FIGURE 34. Connection of indicator to receiver and switch.

Capacitor C_2 is used to stabilize the indicator and prevents the pointer from responding to the low-frequency switching rate. With this type of indicator it is necessary that the receiver furnish an audio-frequency output that in turn is rectified by the rectox unit. A beat-frequency oscillator in the receiver would be desirable to furnish the audio frequency, but due to the inherent instability of the receiver r-f oscillator and many of the transmitter carriers operating at these frequencies, the use of a beat-frequency oscillator is limited. The particular receiver used is equipped with an audio oscillator that modulates the intermediate frequency and furnishes a tone output from a c-w carrier input.

When receiving radar, pulsed at audio frequencies, it is not necessary to use the a-f heterodyne oscillator, providing the repetition rate of the transmitter is sufficient to produce a fair amount of a-f output. The direct current to operate the indicator could have been directly obtained from the a-c line. In this case it would not have been necessary to modulate locally the c-w carrier but it would have been inconvenient in using the receiver for other measurements requiring fixed gain. Although methods of indication were not a part of the problem, there are several others which may be used advantageously with these arrays. These are described below.

4.7 COMPARISON BETWEEN V AND FLAT ARRAYS

The advantages and comparison of the two types of arrays as observed during their development may be summarized as follows:

The V array, using two dipoles (spaced 86 cm) per screen, is electrically a satisfactory unit possessing good directivity and reasonably low polarization errors which may be further improved by careful balance. The directivity is not confined to a narrow sector so that there is little possibility of losing a desired signal located within a known sector of at least 120°.

The construction of the dipoles and transmission line system is such that an accurate balance between the two halves of the array may be readily obtained. This balance can be maintained for long periods of timo without readjustment.

The intersection point of the switched lobes may be chosen by setting the screens to the desired angle, and this point remains reasonably constant throughout the frequency range as the polar patterne are not subject to sudden changes with frequency.

The bearings are sharp, and with interlocking of the lobes at the angle of 17.5° from the lobe maximum, i.e., with the internal angle between screens set at 145°, the bearings are repeatable to approximately ½° throughout the frequency range down to a signal input equal to one-half the receiver noise, measured at a lobe maximum.

There are no reversals in bearing throughout the frequency range and "sense" is therefore unmistakable. If a bearing should be obtained using the back of the screen, i.e., at 180°, it is readily noticed for two reasons:

1. The action of the indicator is reversed.
2. The amplitude of the received signal is greatly reduced.

Mechanically this array appears to be best suited to locations which are semifixed in nature and where space is not of great importance. This is due to the wide turning radius required for the 150- to 300-mc array. If designed for higher frequencies, the array size is proportionately decreased throughout and becomes suitable for portable use.

The large array as used during the tests

proved somewhat awkward to handle in a high wind. This could have been remedied by placing the apex of the screens somewhat ahead of the supporting rotating member, thereby improving the dynamic balance.

The addition of a 300- to 600-mc array attached to the back of the large array, forming a diamond-shaped section as viewed from the top, would also tend to improve the balance and decrease the weather-vane action.

The advantages of the flat array lie in its smaller size, greatly improved rotational balance, and ease of operation.

The bearings are sharp throughout the band and may readily be repeated to better than one-half a degree, a slight improvement existing between the sharpness of this array at certain frequencies and that of the V array.

The average polarization error is slightly lower than that obtained with the V array, and in general, this system appears to be a preferable type for operation on signals which have a reasonable length of transmission period. The reason for this latter qualification is that at some frequencies the lobes are quite sharp, and unless the array is oriented within a few degrees of one of the lobe maxima, the signal may not be picked up. In addition, at a number of frequencies there are reversals of indication that are symmetrically located on either side of the true bearing. The reversals are caused by the way in which the lobes overlap, or in some instances do not fully overlap, a spurious side lobe. In general, a false indication is readily detected either by the amplitude of output, which is relatively weak at the reversal point, or more accurately by reversal of indication. The use of the cathode-ray indicator removes all ambiguity.

Since the polar patterns change rapidly with changes in frequency due to a multiplicity of effects resulting from phase shift caused by the electrical changes in spacings and transmission-line linkage, it is not possible to locate the optimum cross-over point of lobe intersections at more than a few frequencies. At the remaining frequencies the intersections fall where they may, although with the antenna spacings and lines cut to the dimensions shown, the performance approaches the maximum obtainable over the frequency range and does not depart greatly from the optimum or desired performance except at the higher end of the frequency range, where the intersection of the lobe drops to an amplitude somewhat below that desirable for optimum signal-to-noise ratio. The effect on bearing sensitivity in this case is to increase the angular sensitivity to the detriment of the radio frequency sensitivity as indicated by the calculations for optimum performance.

In either type of array the use of two coaxial shielded flexible cables appears to have an advantage over the more commonly used spaced air-dielectric twin pairs. This is apparent electrically from the good degree of balance that may be obtained by simply cutting to the same mechanical length leads that are to be matched. The uniformity of cable, of reliable manufacture, obtained from the same reel, is sufficient in most cases for one to be reasonably sure of better than passable matching. This was electrically measured and checked several times in the course of changes and development. It is believed that such r-f cable is adaptable to feeders for the elevated H Adcock-type antenna, where strict symmetry and balance are required.

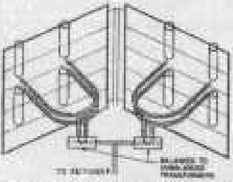
4.8 COMPARISON BETWEEN DIFFERENTIALLY CONNECTED SCREEN ARRAYS AND H ADCOCKS

Some data were obtained using the V-1 array differentially connected, and the V-2 array with each pair differentially connected to the opposite pair, the dipoles of each reflector being connected in phase. The latter connection is indicated in Figure 35. This construction is interesting because, while it resembles an H Adcock, the screen angle is such that the spurious side lobes normally obtained with multiple dipoles on a flat array are absent due to the fact that at wide angles from the null, one antenna is shielded by the screen from the signal source and hence the system no longer acts as a differential system.

It is difficult to make quantitative comparison between this differentially connected screen array and the more common elevated H Adcock which it resembles without a side-by-side check using field-intensity equipment. However, it is

possible to indicate certain generalities and limitations.

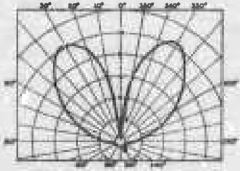
The type of Adcock to be considered as a reference is of the balanced elevated H design most commonly used on these frequencies. The selection of dipole length and spacing would



[illegible] V-2 array.

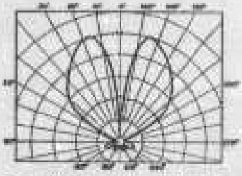
vary slightly with the designer's choire, but a normal unit would have a dipole spacing such that, at the minimum wavelength, the spacing would not exceed $\lambda/2$. In any case, the choice of spacing would be such that the polar patterns would not divide into more than two lobes. Four lobes, such as would appear at a spacing of λ would, of course, be unusable as there would be no rapid way to distinguish which of the four bilateral minima would be correct. Therefore the conventional spacing would be such as to give a pattern approaching a cosine curve, and might be in the order of $\lambda/6$ to $\lambda/2$, giving at the smaller spacing a maximum response equivalent to that of a single dipole in free space, and at the greater spacing a maximum response double this value.

If, however, single dipoles of length equal to the above Adcock but of suitable diameter are arranged at an appropriate distance in front of a V screen and these dipoles are differentially connected, there are three important results. First, the gain of the dipoles is increased in a direction normal to the screen by a factor of approximately 5 db; second, the response pattern becomes unidirectional; third, the presence of surrounding objects outside of the field of the lobes does not materially affect the bearings, and therefore the effect of reradiating objects, located behind the screen, can be tolerated to an increased degree.



[illegible] V-2 array, [illegible]

It becomes possible to utilize a spacing of greater than one wavelength between dipoles placed in front of an angle screen and still obtain only two lobes. There are no lobes on the back of the screen, and hence, no "null" or balance between lobes in a direction parallel to



[illegible] V-2 array, [illegible]

the plane of the screen. Minima exist along the plane of the screen, but these cannot be confused with a null as rotation beyond the line parallel with the screen does not increase the output. The advantage of the increased dipole separation up to one wavelength or more is that the angular sensitivity is increased.

The polar patterns resulting when the V array, using two dipoles per screen, is connected as a balanced system using a differential connection between the two pairs are indicated in Figures 36 and 37. The spacing between dipoles in this case was 1.22λ between the midpoint of each pair at the highest frequency, 300 mc. The angle between screens was not optimum for this use, but the polar patterns illustrate at three frequencies the forward gain and indicate the extreme sharpness of nulls.

The forward gain of each pair of two dipoles in front of a screen, at an angle normal to the screen, is approximately 8 db over the two-to-one frequency band when compared to a single dipole in free space, $\lambda/2$ long at each frequency of comparison. The gain measurements for the pair of antennas in front of a screen are given in Figure 22.

The measured polarization errors are low and are given in Table 1. The tilted-dipole method was used in measuring these errors, and this may be roughly correlated with measurements made with the variable-phase polarization transmitter by reference to the measurements made on the lobe-switched V-1 array where both methods were used. In general, it appeared that the tilted-dipole method was quite satisfactory at these frequencies, particularly when the errors were low. When properly used it is indicative of the general performance to be expected.

The V-2 array mentioned above is that discussed in preceding sections as the V-2 array, wherein it was connected as a lobe-switched device. The edges of the screens were insulated from the supporting pole. Separate balanced-to-unbalanced transformers were used at the back of each screen and grounding to the support pole was made through the shield of the coaxial cable leading from the transformers to the central support shaft.

DESIGN OF BALANCED-TO-UNBALANCED TRANSFORMERS

The transformers used in both the V and flat arrays for converting the balanced dipole systems to an unbalanced line are designed along the same lines. Each consists of two short-circuited coaxial sections, $\lambda/4$ long at the mean frequency, one connected across each half of the balanced line. Where reactance cancellation is desired, a shorted half-wave section may be inserted in series with the grounded side of the unbalanced line. The circuit is shown schematically in Figure 38.

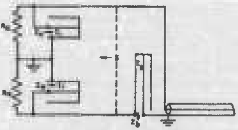


FIGURE 38. Schematic of transformers for connecting balanced doublets to unbalanced lines.

If the dipole impedance is taken to be resistive, each half may be represented by R_a. The $\lambda/4$ lines have a characteristic impedance Z_1 and input impedance Z_a, while the corresponding impedances for the half-wave line are Z_2 and Z_b. The impedance looking toward the dipoles at Z is

$$Z = \frac{2Z_a R_a}{Z_a + R_a}. \qquad (1)$$

For lossless lines, $Z_a = jZ_1 \tan\phi$, ϕ being the phase length.

Then
$$Z = \frac{2jZ_1 \tan\phi\, R_a}{jZ_1 \tan\phi + R_a}. \qquad (2)$$

$$Z = \frac{2Z_1^2 \tan^2\phi\, R_a}{Z_1^2 \tan^2\phi + R_a^2} + j\frac{2Z_1 \tan\phi\, R_a^2}{Z_1^2 \tan^2\phi + R_a^2} \qquad (3)$$

$$= \frac{2R_a}{1 + \dfrac{R_a^2}{Z_1^2 \tan^2\phi}} + j\frac{2Z_1 \tan\phi}{1 + \dfrac{Z_1^2 \tan^2\phi}{R_a^2}}. \qquad (4)$$

If Z_1 is made greater than R_a, and ϕ is in the vicinity of 90°:

$$Z_1^2 \tan^2\phi \gg R_a^2. \qquad (5)$$

Then
$$1 + \frac{R_a^2}{Z_1^2 \tan^2\phi} \approx 1, \text{ and}$$

$$1 + \frac{Z_1^2 \tan^2\phi}{R_a^2} \approx \frac{Z_1^2 \tan^2\phi}{R_a^2} \qquad (6)$$

hence
$$Z = 2R_a + j\frac{2R_a^2}{Z_1 \tan\phi}. \tag{7}$$

Also
$$Z_b = jZ_2 \tan 2\phi \tag{8}$$

and for reactance cancellation—

$$jZ_2 \tan 2\phi + \mathrm{Im}\,(Z) = 0 \tag{9}$$

$$\therefore\ Z_2 \tan 2\phi = -\frac{2R_a^2}{Z_1 \tan\phi} \tag{10}$$

or
$$2R_a^2 = -Z_1Z_2 \tan\phi \tan 2\phi \tag{11}$$

$$= -Z_1Z_2 \frac{2\tan^2\phi}{1-\tan^2\phi}. \tag{12}$$

Again, for ϕ in the vicinity of 90°,

$$\frac{2\tan^2\phi}{1-\tan^2\phi} \simeq -2 \tag{13}$$

so that $R_a^2 = Z_1 Z_2$. If this condition is satisfied, the residual series reactance introduced by the transformer is minimized through a frequency range over which the approximations made are valid.

The error due to the approximation

$$Z_1^2 \tan^2\phi \gg R_a^2 \tag{14}$$

may be made negligible by making Z_1 much greater than R_a. The practical limitation is the large ratio of diameters required in the coaxial elements for high Z_1; also, Z_1 increases much more slowly than this ratio (as the logarithm of the ratio).

The other approximation:

$$\frac{\tan^2\phi}{\tan^2\phi - 1} = 1 \tag{15}$$

when made over an effective phase length of 60° to 120°, corresponding to a two-to-one frequency range, introduces an error of 33⅓ per cent at the two extremes, and, if desired, may be taken into account.

If an open-circuited quarter-wave line is substituted for the half-wave line, the condition for cancellation becomes

$$Z_2 \cot\phi = \frac{2R_a^2}{Z_1 \tan\phi} \tag{16}$$

or
$$2R_a^2 = Z_1Z_2 \tan\phi \cot\phi$$
$$= Z_1Z_2. \tag{17}$$

This is exact, and eliminates the second approximation, but requires twice the previous value for Z_1Z_2. The shorted half-wave section also assists in keeping current from traveling down the outer conductor of the coaxial down lead and may be more desirable for direction-finding use, where stray fields must be kept to a minimum.

The transformers may be seen in Figure 4 mounted on the back of the V-1 array. The mechanical arrangement of the transformer is shown in Figure 39.

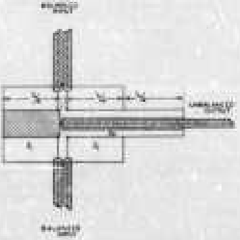


Figure 39. Mechanical arrangement of transformers.

A transformer of this type is the equivalent of an electrostatically screened transformer, and prevents a balanced system from acting as a grounded antenna, that is, it prevents a dipole from responding as a unit to a potential gradient between it and the ground. It enables the system to be balanced by merely establishing balance from the transformer to the dipole; the unbalanced line from the transformer to the receiving equipment does not, of course, require such treatment.

The reactance cancellation line may be omitted under certain circumstances. If the resultant impedance of the antenna circuits, as seen at the transformer, is low, the transformer characteristic impedance may be made sufficiently high with reasonable diameters of the quarter-wave elements so that the residual series reactance introduced may be negligibly

small. Compensation under these circumstances would hardly be justified, in view of the fact that the antenna impedance is itself partially reactive, and may contribute an appreciably larger reactive component than the transformer.

4.10 DETERMINATION OF GROUND CONSTANTS

To obtain the magnitude of ground reflection effects required for the correlation of polarization error data, a number of methods of measurements were reviewed in the literature. The normal incidence method developed by McPetrie[4] appeared to be the most likely to yield accurate results. Essentially it consists of setting up a field from an elevated source, and sampling the standing wave pattern set up by the direct and ground-reflected waves at normal incidence. The ground constants may be deduced from the data so obtained.

Primarily because of the special setup required for this method, a comparatively simple laboratory method was developed, more suitable for the available facilities. The results obtained show good agreement with published data on the ground constants in the vicinity of the test site, as well as with oblique incidence field-intensity measurements made during polarization error investigations. The degree of correlation may be observed in Figures 14 and 15, where the measured standing wave pattern is shown with the pattern calculated on the basis of the measured ground constants.

The method employs a short section of coaxial line as an extension to a slotted coaxial measuring line, both having the same transverse dimensions, and consequently the same characteristic impedance with air as dielectric. Provision is made for either open or short circuiting the end of the extension. The input impedance of the extension is measured, when a sample of the ground in question is substituted for the air as dielectric, for the two conditions. The impedances so obtained may be represented as follows:

$$Z_{oc} = |Z_{oc}| \underline{/\theta_{oc}} \qquad (18)$$

$$Z_{sc} = |Z_{sc}| \underline{/\theta_{sc}} \qquad (19)$$

The characteristic impedance of the extension is then:

$$Z_{s}{}^{2}{}_{\text{ground}} = Z_{oc} Z_{sc}$$

$$= |Z_{oc}|\,|Z_{sc}| \underline{/\theta_{oc} + \theta_{sc}}$$

$$= |Z_{oc}Z_{sc}| \left[\cos(\theta_{oc} + \theta_{sc}) + j \sin(\theta_{oc} + \theta_{sc})\right] \qquad (20)$$

$$= r + jx \qquad (21)$$

where Z_{oc} = open-circuit impedance,

Z_{sc} = short-circuited impedance.

Assume a harmonic plane wave propagated longitudinally along the coaxial line. The field components are transverse; E_r is the radial electric field and H_ϕ the tangential magnetic field, as in Figure 40. The other field components are zero.

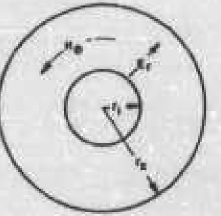


FIGURE 40. Electric and magnetic fields in coaxial line.

The ratio of the electric to the magnetic field of a plane wave at a point is the intrinsic impedance of the medium for plane waves:

$$\frac{E_r}{H_\phi} = \mathbf{Z} = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\epsilon}} \qquad (22)$$

Here μ, ϵ, σ, are the permeability, permittivity, and conductivity, respectively, and ω the angular velocity.

To obtain the relation between the intrinsic impedance of the dielectric and the characteristic impedance of the line, the longitudinal current and the transverse voltage are required.

The longitudinal current is

$$I_L = \oint \mathbf{H} \cdot d\mathbf{s}. \tag{23}$$

Because of circular symmetry, H_θ is independent of θ; then, along a circle of radius r, since H_θ lies along the circle,

$$I_L = \oint H_\theta ds. \tag{24}$$

Since H_θ is constant for a given r,

$$I_L = H_\theta \oint ds \tag{25}$$

$$= H_\theta(2\pi r). \tag{26}$$

If the conductivity of the inner and outer cylinders is much greater than that of the dielectric, the current enclosed within the path of integration $(r_1 < r < r_2)$ may be assumed to flow entirely on the inner conductor. Equation (26) therefore gives the longitudinal current on the inner conductor.

The transverse voltage between the outer and inner conductors is defined as the line integral of the electric field between the conductors along a path lying in a transverse plane. A radial path is most convenient:

$$V_T = \int_r \mathbf{E} \cdot d\mathbf{s} \tag{27}$$

Since E and ds are both directed along a radius, it follows that

$$V_T = \int_{r_1}^{r_2} E_r dr \tag{28}$$

from (22)
$$E_r = H_\theta \mathbf{Z} \tag{29}$$

from (26)
$$H_\theta = \frac{I_L}{2\pi r} \tag{30}$$

substituting (30) in (29)

$$E_r = \frac{I_L}{2\pi r} \mathbf{Z} \tag{31}$$

substituting (31) in (28)

$$V_T = \int_{r_1}^{r_2} \frac{I_L}{2\pi r} \mathbf{Z} dr \tag{32}$$

$$= \frac{I_L \mathbf{Z}}{2\pi} \Big| \log r \Big|_{r_1}^{r_2} \tag{33}$$

$$= \frac{I_L \mathbf{Z}}{2\pi} \log \frac{r_2}{r_1}. \tag{34}$$

The characteristic impedance of a line is defined as the ratio of the tranverse voltage to the longitudinal current. Hence

$$\frac{V_T}{I_L} = Z_c = \mathbf{Z} \frac{\log \frac{r_2}{r_1}}{2\pi}. \tag{35}$$

This is the required relation connecting Z_c and $\mathbf{Z}$. The characteristic impedance of a coaxial line is thus given by the product of a geometrical factor and the intrinsic impedance of the dielectric medium. If two dielectrics, air and ground, are compared in a line of fixed geometry,

$$\frac{Z_{c\,(\text{ground})}}{Z_{c\,(\text{air})}} = \frac{\mathbf{Z}_{(\text{ground})}}{\mathbf{Z}_{(\text{air})}} \tag{36}$$

$$= \frac{\sqrt{\dfrac{j\omega\mu_g}{\sigma_g + j\omega\epsilon_g}}}{\sqrt{\dfrac{j\omega\mu_0}{\sigma_0 + j\omega\epsilon_0}}}. \tag{37}$$

Here the subscript g refers to ground used as dielectric, and 0 to air dielectric.

or
$$\frac{Z_{c\,(\text{ground})}}{Z_{c\,(\text{air})}} = \frac{\sqrt{\dfrac{\mu_g}{\epsilon_g - j\dfrac{\sigma_g}{\omega}}}}{\sqrt{\dfrac{\mu_0}{\epsilon_0 - j\dfrac{\sigma_0}{\omega}}}}. \tag{38}$$

If the permeability of the ground is taken to be equal to that of air or free space,

$$\frac{Z_{c\,(\text{ground})}}{Z_{c\,(\text{air})}} = \frac{\sqrt{\epsilon_0 - j\dfrac{\sigma_0}{\omega}}}{\sqrt{\epsilon_g - j\dfrac{\sigma_g}{\omega}}}. \tag{39}$$

For free space, $\sigma = 0$,

$$\frac{Z_{c\,(\text{ground})}}{Z_{c\,(\text{air})}} = \frac{\sqrt{\epsilon_0}}{\sqrt{\epsilon_g - j\dfrac{\sigma_g}{\omega}}}. \tag{40}$$

Thus far mks units have been used. The equation for converting ϵ to electrostatic units is

$$\epsilon_{\text{mks}} = \frac{\epsilon_{\text{esu}}}{4\pi \times 10^{-11} c^2}. \tag{41}$$

For converting σ to electromagnetic units the following equation applies:

$$\sigma_{\text{mks}} = 10^{11} \sigma_{\text{emu}}. \tag{42}$$

Making these conversions, noting that $\epsilon_0 = 1$ in esu and dropping the subscripts g and 0,

$$\sqrt{\epsilon_{\text{soil}} - j\frac{18 \cdot 10^{14}\,\sigma_{\text{soil}}}{f_{mc}}} = \frac{Z_{c\,(\text{air})}}{Z_{c\,(\text{ground})}}\,. \tag{43}$$

The quantity under the radical is known as the complex dielectric constant, and may be represented as $\epsilon' - j\epsilon''$.

$$\epsilon' - j\epsilon'' = \frac{Z_c{}^2{}_{(\text{air})}}{Z_c{}^2{}_{(\text{ground})}}\,, \tag{44}$$

and from equation (21)

$$\epsilon' - j\epsilon'' = \frac{Z_c{}^2{}_{(\text{air})}}{r + jx} \tag{45}$$

$$= \frac{rZ_c{}^2{}_{(\text{air})}}{r^2 + x^2} - j\frac{xZ_c{}^2{}_{(\text{air})}}{r^2 + x^2} \tag{46}$$

where the values of r and x are to be obtained from equation (20).

Equating the real and imaginary parts:

$$\epsilon' = \frac{rZ_c{}^2{}_{(\text{air})}}{r^2 + x^2}\,. \tag{47}$$

$$\epsilon'' = \frac{xZ_c{}^2{}_{(\text{air})}}{r^2 + x^2}\,. \tag{48}$$

Using this method, ϵ' was found to be equal to 10, and $\sigma = 8.8 \times 10^{-14}$ emu for the ground at the test site.

Care must be exercised in packing the earth into the line extension to maintain the same density in the actual and measuring conditions. Repeated measurements indicated practically constant σ, while ϵ' showed some variation depending on the moisture content of the earth. The value of 10 may be taken as representing average conditions.

4.11 IMPEDANCE OF A CYLINDRICAL DIPOLE BEFORE A REFLECTOR

As indicated above, considerable variations were encountered between the measured impedance characteristics of the dipoles used on the V-1 screen and the theoretical characteristics based on prolate spheroidal dipoles as given by Stratton and Chu.[3] Certain other treatments of the problem were examined in an attempt to obtain better agreement between experimental data and existing theory.

The values of self impedance obtained from Hallen's formulas as given by King and Blake,[6] and King and Harrison, and the values we calculated from the formulas of Schelkunoff,[8,9] were compared to the experimental values of impedance obtained on the V-1 array. The latter is the impedance in the presence of the reflector; corrections for the mutual impedance between the dipole and its image were to be applied on the basis of the theory developed by Brown.[10]

The values of self resistance based on Hallen's formula were found to be too high; results obtained from Schelkunoff's formula showed better agreement, but not good enough for engineering purposes; a correction[11] for the concentrated capacitance in the vicinity of the gap brought this theory into much closer agreement with the measurements. It may be mentioned that a modification of Hallen's solution by Gray,[12] yields better results than the original, but not as good as Schelkunoff's. A comparison of the latter with our measurements is made below.

The self impedance of a cylindrical dipole is given by:

$$Z_{\text{self}} = \frac{Z_a R_t \sin\beta l + j\{[X_t - f_2(2\beta l)]\sin\beta l - [Z_a - f_1(2\beta l)]\cos\beta l\}}{[Z_a + f_1(2\beta l)]\sin\beta l + [X_t + f_2(2\beta l)]\cos\beta l - jR_t\cos\beta l} \tag{49}$$

where

Z_{self} = self impedance

R_t = terminal resistance

$= 60\ \text{Cin}\ 2\beta l + 30(C + \ln\beta l - 2\ \text{Ci}\ 2\beta l + \text{Ci}\ 4\beta l)\cos 2\beta l + 30\,(\text{Si}\ 4\beta l - 2\ \text{Si}\ 2\beta l)\sin 2\beta l$

X_t = terminal reactance

$= 60\ \text{Si}\ 2\beta l + 30\,(\text{Ci}\ 4\beta l - \ln\beta l - C)\sin 2\beta l - 30\ \text{Si}\ 4\beta l\cos 2\beta l$

$$f_1(2\beta l) = 60(\text{Cin}\ 2\beta l - 2\sin^2\beta l)$$

$$f_2(2\beta l) = 60(\text{Si}\ 2\beta l - \sin 2\beta l)$$

$$Z_a = 120\ln\frac{2l}{a} - 120$$

$$\beta = \frac{2\pi}{\lambda}$$

λ = wavelength

l = half length of dipole

a = radius of dipole

Ci( ) = cosine integral function, tabulated[13,14]

Si( ) = sine integral function, tabulated[13,14]

Cin( ) = C + ln ( ) − Ci ( )

C = Euler's constant (= 0.5772)

The resistive component of the self impedance is:

$$R_{self} = \frac{Z_0}{D} R_1 [Z_0 + f_2(2\beta l) \sin 2\beta l - f_1(2\beta l) \cos 2\beta l] \tag{50}$$

The reactive component of the self impedance is:

$$X_{self} = \frac{Z_0}{D} \{ \tfrac{1}{2} [R_t^2 + X_t^2 - Z_0^2 + f_1^2(2\beta l) + f_2^2(2\beta l)] \sin 2\beta l + [f_1(2\beta l) f_2(2\beta l) - Z_0 X_1] \cdot \cos 2\beta l + [X_t f_1(2\beta l) - Z_0 f_2(2\beta l)] \} \tag{51}$$

where

$$D = (R_1 \cos \beta l)^2 + \{ [Z_0 + f_1(2\beta l)] \sin \beta l + [X_t + f_2(2\beta l)] \cos \beta l \}^2 \tag{52}$$

Figure 41, curve a, is a plot of the resistive component of the free space input impedance of the dipole used in the V-1 array, as obtained

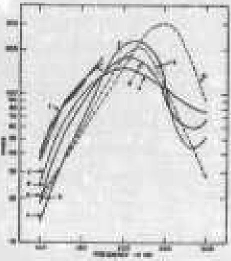


FIGURE 41. Resistance characteristics of dipole in front of reflector.

from equation (50). The resistance corrected for a gap capacitance of 2.0 $\mu\mu f$ is shown in curve b, while values of the measured resistance are given in curves c, d, and e for spacings from the reflector of 33, 28.5, and 24.2 cm respectively. Curve b is in good qualitative agreement with the measurements. The outstanding differences are the downward displacement along the frequency scale of the experimental curves, and the relatively high maximum value of the theoretical curve. Better agreement is possible if a decrease in velocity of propagation greater than predicted by the theory is assumed.

The corrections for mutual impedance were not applied to the theoretical curves, as the discrepancies between the latter and the experimental curves are of the same order of magnitude as the corrections involved. To test the applicability of the mutual impedance theory to dipoles of the proportions used, the reverse process was adopted. Starting with the three measured resistance curves of Figure 41, the three corresponding self-resistance curves were deduced by means of the inverse corrections for mutual resistance. The three self-resistance curves so obtained are almost identical up to a full-wave dipole length, indicating that the theory is applicable up to this limit. The group of three self-resistance curves is identified as f on Figure 41.

The mutual resistance is accounted for in the following manner: The resistive component of the coefficient of radiation coupling is known to be independent of dipole length for two identical parallel nonstaggered thin dipoles, up to one wavelength long. It may be defined as the ratio of the resistive component of mutual impedance to the resistive component of self impedance. Thus, if the coefficient is known, either the self or mutual resistance may be obtained provided one or the other is known.

The mutual resistance between two dipoles as limited above is:[15]

$$R_{mutual} = \frac{30}{\sin^2 \beta l} \{ 2(2 + \cos 2\beta l) \mathrm{Ci}\, 2\beta s - 4 \cos^2 \beta l (\mathrm{Ci}\, \beta E + \mathrm{Ci}\, \beta F) + \cos 2\beta l (\mathrm{Ci}\, \beta G + \mathrm{Ci}\, \beta H) + \sin 2\beta l (\mathrm{Si}\, \beta H - \mathrm{Si}\, \beta G - 2\, \mathrm{Si}\, \beta F + 2\, \mathrm{Si}\, \beta E) \} \tag{53}$$

The notation here is the same as above, with the additions

s = half the distance between the two dipoles (or the distance from one dipole to a reflector)

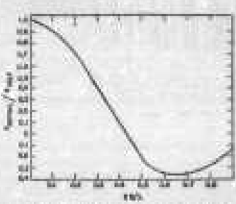
$$E = (\sqrt{4s^2 + l^2} - 1)$$
$$F = (\sqrt{4s^2 + l^2} + 1)$$
$$G = (2\sqrt{s^2 + l^2} - 1)$$
$$H = (2\sqrt{s^2 + l^2} + 1)$$

Since equation (53) is the asymptotic expression for the mutual resistance of two infinitely thin dipoles, it may not be compared with equation (50) directly to obtain the resistive component of the coefficient of coupling.

The following expression may be used:

$$R_{self} = 30\{(1 - \cot^2 \beta l)\ (\text{Cin}\ 4\beta l) + 4 \cot^2 \beta l\ (\text{Cin}\ 2\beta l) + 2 \cot \beta l\ (\text{Si}\ 4\beta l - 2\ \text{Si}\ 2\beta l)\}\ . \quad (54)$$

The ratio of equation (53) to equation (54) is the resistive component of the coefficient of radiation coupling. A plot of R_{mutual}/R_{self} is given in Figure 42 as a function of $2s/\lambda$.



[illegible]

The deduced values of self resistance given by curve f of Figure 41 were obtained using Figure 42, since the input resistance may be expressed as:

$$R_{in} = R_{self} - R_{mutual}\ . \quad (55)$$

Calculation of the self reactance is based on equation (51). The mutual reactance is accounted for as follows: the phase angle of the mutual impedance between two identical parallel nonstaggered thin dipoles, up to one wavelength long, is to a first approximation independent of the length, and a linear function of the spacing. The linear connection is, for s greater than 0.1λ,

$$\phi = -312\frac{2s}{\lambda} + 42 \quad (56)$$

Here ϕ is the phase angle in degrees; the other symbols are as previously used.

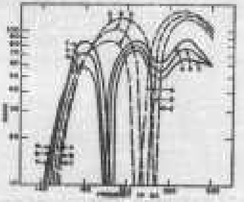
From the relation

$$\tan \phi = \frac{X_{mutual}}{R_{mutual}} \quad (57)$$

and the previously obtained values of R_{mutual}, X_{mutual} may be determined. The total input reactance is then

$$X_{in} = X_{self} - X_{mutual}\ . \quad (58)$$

Since the expression for X_{mutual} contains $\tan \phi$ as a factor, the correction for X_{mutual} as expressed in equation (58) may have much larger values than the corresponding correction R_{mutual} in equation (55). The former correction was therefore applied directly to the values of X_{self} as obtained from equation (51), for



[illegible]

the dipoles used on the V-1 array, for three values of the spacing s from the reflector. Figure 43 is a plot of these: curve A is for a spac-

ing of 33 cm, or $\lambda/4$ at 227 mc; curve B, 28.5 cm, 263 mc; curve C, 24.2 cm, 310 mc. The corresponding experimental curves are shown at a, b, and c of the same figure. An examination of these curves indicates that the corrections for mutual reactance are of the correct order of magnitude. As in the case of the resistance curves of Figure 41, the experimental reactance curves correspond to a velocity of propagation lower than that predicted by the theory, and the values of computed reactance are high.

The accuracy of wavelength determinations made in the course of impedance measurements is sufficiently high to preclude the possibility of experimental error accounting for the difference in velocity of propagation indicated by these two sets of curves.

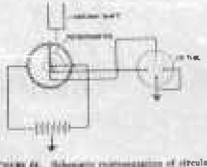
4.8 CATHODE-RAY INDICATION AND AUTOMATIC CONTROL

Subsequent to the expiration of the contract. several methods of cathode-ray indication equivalent to *plan position indicators* [PPI] were developed and an automatic control was added to the flat array to indicate the practicability of the arrays when used for direction finding.

In searching it is desirable to rotate the antenna array and provide a visual means of locating the azimuth. To accomplish the rotation and also to provide means of automatically obtaining a bearing once the signal quadrant is known, an amplidyne servo system was installed. This was used to drive the antenna shaft either (1) through means of a manually operated selsyn control, or (2) automatically through suitable output amplifiers connected to the differential voltage developed across the indicator meter circuit. These two arrangements provided means for rotating the array to any desired azimuth when the selsyn was used, or to automatically orient the array to the signal bearing when the receiver output differential voltage was used as the control. The maximum speed of antenna rotation from either arrangement was 6 rpm.

In addition to the L-R indicator meter, which indicates when the array is on bearing, a long persistent CR tube was used in the combinations which follow. The means of placing the CR spot or trace, depending upon the presentation employed, was to gear a resistor control to the antenna shaft and provide electrical connections from this to the deflecting plates of the CR tube. The resistor consists of a circular strip with two brushes at 90° from each other. If direct current is applied to the proper terminals of the resistor strip, the CR spot is moved from the center of the tube to an angular position corresponding to the location of the resistor brushes.

Under the above condition, rotation of the brushes produces a circular trace. The resistor control being geared to the antenna shaft, therefore, produces a trace which is synchronized with the antenna array. This is shown schematically in Figure 44. Several forms of



[illegible] circular- [illegible]

presentation were tested, which, in each case, indicated the array position and the relative amplitude of the signal.

INDICATION PRESENTATION

The first method employed was to superimpose on the circular trace the differential voltage developed across the L-R indicator meter. The pattern, Figure 45A, is such that signals to the left of the bearing appear as an increase in the circle and are, therefore, outward, while at the right of the bearing the patterns are inward. At the bearing position, the trace is evenly divided in amplitude about the circle. This arrangement is unmistakable but also unsymmetrical and, therefore, requires a slight amount of interpretation.

A second method is to connect the d-c voltage across the rotatable resistor and in series with the output rectifier from the receiver without going through the L-R meter switch. In this line at the bearing is obtained as shown in Figure 45B.

Another arrangement is to drive the circular trace inward rather than outward. This forms

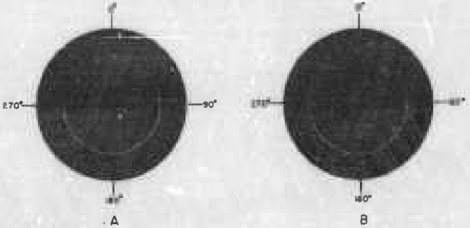


FIGURE 46. A shows pattern secured by superimposing on circular trace differential voltage developed across the L-R indicator meter. B shows d-c voltage connected across rotatable resistor and in series with rectifier output from receiver without going through L-R meter switch.

case the circular trace is maintained and a pattern which increases the circular trace on or to either side of the null and drops to a balanced a more suitable pattern, since the bearing is indicated by an arrow formed by the parts of the face of the tube which were not illuminated

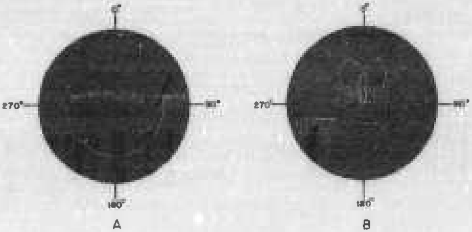


FIGURE 46. A shows circular trace driven inward rather than outward. B shows lobe-switched output connected to produce trace of two lobes.

by the trace. Figure 46A shows this pattern.

A fourth arrangement is to connect the lobe-switched output in such a manner as to produce the trace of both antenna lobes. In this case the intersection of the lobes indicates the bearing as illustrated in Figure 46B.

Other arrangements were employed connecting the antenna array as a differential array forming, in effect, an Adcock antenna and tracing the pattern and null directly on the tube. (See Figure 47A.) A reversed connection of this arrangement, shown in Figure 47B,

It appears that the maximum utility of the CR tube indicator is to locate roughly the source of the signal with an accuracy of $\pm 2°$. A bearing may be read more accurately if obtained by the automatic control once the quadrant has been located. The bearing scale for the automatic control was read directly from the azimuth scale mounted on the antenna array, although provisions are made in the amplidyne system to read the indication from a separate selsyn which is geared to the antenna shaft.

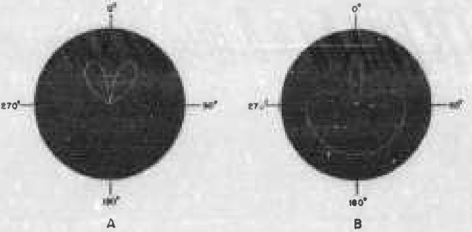


FIGURE 47. A shows effect of antenna connected as differential array forming Adcock antenna. B shows effect of reversing connections from those producing A.

produces a trace which draws a line outward to the edge of the CR tube at the bearing indication point. The antenna arrangements for the two latter patterns do not require the lobe-switching mechanism and are, therefore, somewhat simpler. However, this arrangement cannot readily be employed as an automatic direction finder or be electrically connected to the servo system so that the bearing is obtained automatically.

Many other presentation arrangements are possible using the CR tube. The methods of indication presentation suggested above, with the exception of the system which presents the direct or reversed patterns of the Adcock arrangement, are based on the lobe switching of the antennas. It should be noted that all of the above illustrations show the bearing at 0° azimuth. The patterns in each case rotate with bearing.

Chapter 5

ERRORS IN DIRECTION FINDERS

NUMEROUS PROJECTS under Division 13 were concerned with the fact that direction finders of various types do not give consistent or accurate bearings in spite of the fact that they can be erected with great care and made up of precision apparatus. Some of these errors were found to be due to the fact that elevated structures do not have all parts equidistant from the reflecting or semi-conducting ground; that waves arriving from the ionosphere are polarized in heterogeneous ways; that waves traversing regions near the magnetic poles do not always follow the great-circle route; and there are still other reasons why d-f results do not have the accuracy desired. The background for these troubles will be found discussed in Chapters 1 and 2, and, in fact, throughout the summaries of d-f projects reported in this volume.

5.1 PROJECT C-17[a]

This was the [illegible] of a continuing series of projects for studying certain errors of shielded-U Adcock direction finders.

Under Project C-17[1] will be found a general review of the directive properties of radio waves and wave collectors, giving reasons why simple loop and dipole antenna d-f systems are not accurate under normal conditions of h-f wave propagation. The fact is that spaced-antenna systems to eliminate the faults of the simple loop or dipole are in theory highly accurate but in practice are not so. The importance of taking rapid bearings, of making all antennas of a given spaced-antenna d-f system identical, and of limiting unwanted pickup from extraneous conductors is discussed in this review which also evaluates various known wave collecting systems.

This review found the shielded-U Adcock especially promising. Since the nature and extent of the shield required to produce sufficiently accurate bearings on sky waves had never been fully studied, Project C-17 was set up to study the design and properties of this particular type of antenna system. Attention was directed particularly toward portable equipment.

A precise, demountable, shielded-U Adcock antenna was built on top of a station wagon, for portability, and a receiver with calibrated attenuator was installed in the station wagon to measure antenna responses under various conditions of wave incidence. Conductors were provided for building up elevated artificial ground planes of varying extent and completeness. A local source of test signals of definite polarization was provided, together with a balloon and rigging to elevate this source for production of sky-wave signals. A transit was used for observing polarization and direction of arrival of the test signals.

Measurements were made with this system over the range 7 to 18 mc with antennas connected to the input transformer of the receiver directly or through cathode followers, the two methods giving about the same errors. The quantities measured were mostly maximum/minimum ratios for directive patterns and minima positions for ground waves, and the ratio of maximum responses to vertically and horizontally polarized ground and sky waves.

Trouble was experienced from the beginning with the inadequacy of the artificial ground systems tried as a part of the shielding of the Adcock U—trouble which has been observed in all other d-f projects summarized in this volume. Radial-wire counterpoises were found to be wholly inadequate, radial plus ring-wire counterpoises gave good results on ground waves but had excessive errors on unfavorably polarized sky waves. Netting with radial wire extensions, shown undergoing tests in Figure 1, worked fairly well with ground waves and showed only moderately excessive errors with

[a] Contract No. NDCrc-149, Radio Corporation of America.

sky waves. The netting, however, was not conveniently portable.

The general conclusion was that a carefully made portable shielded-U Adcock using a demountable elevated counterpoise can be highly accurate for horizontally arriving signals, but cannot be made outstandingly free of polarization errors on downcoming signals without undue sacrifice of portability. Standard-wave errors of the order of 10° at best were attained.

3.1.1 Apparatus Employed

A continuous copper disk 8 ft in diameter was mounted on top of the station wagon and determined the size of the antenna system. Thus the antenna spacing was arbitrarily set at two-thirds of the disk diameter or about 5½ ft. This spacing was $\lambda/6$ at 30 mc and was $\lambda/36$ at 5 mc. This small spacing made the system rather insensitive, a 1° change in azimuth of an arriving signal producing a phase change of antenna voltage of only 10 minutes of arc at 5 mc. The height of the antennas was 12½ ft, which was $\frac{3}{8}\lambda$ at 30 mc.

Many types of counterpoise systems were investigated and the one with the greatest density of conductors was best, but a larger one with fewer conductors was more practical and fairly good. This was made up of 48 radial wires each 100 ft long attached to the outer perimeter of a 65-ft diameter spider web arrangement with 24 radial and 8 ring wires, with the 8-ft copper disk in the center.

Considerable work was carried out with balloons, with attendant difficulties which limited the amount of downcoming-wave data obtained.

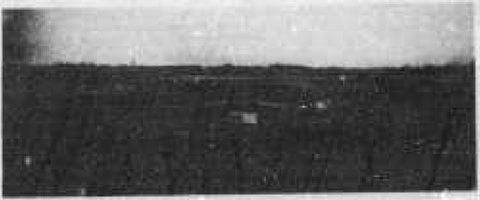


FIGURE 1. Artificial ground system composed of netting with radial wire extensions, shown undergoing tests.

In constructing the test oscillator to be used in the work with the shielded-U Adcock, care was taken to see that the purity of polarization was high. This was secured by making the test oscillator long and narrow to minimize the possibility of r-f current flow in any direction other than that of the antenna rods attached to its ends. Electrical symmetry was provided by connecting the case of the miniature battery-powered transmitter to the center of the coil feeding the two rods of the symmetrical dipole antenna.

3.2 PROJECT C-38[b]

In earlier work, tests had been made of a counterpoise made up of radial wires and ring wires connected at the points where rings crossed radials. Further tests were made under C-38[b] with a counterpoise of ring wires only. Results were, as expected, decidedly worse than with counterpoise arrangements tried earlier.

[b] Contract No. OEMsr-338, Radio Corporation of America.

The purity of polarization of the test transmitter developed under Project C-17 was examined and it was found that the ratio of vertical receiving antenna output with transmitter antenna horizontal and then vertical was over 500.

Some unsuccessful trials were made of a large kite to supplement the balloon as support for a source of high-angle downcoming waves. The balloon rigging was revised to give improved operation over a wider range of conditions.

8.8.1 Tests at Holmdel

Arrangements were made to take the balloon rigging, test transmitter and other auxiliary apparatus to Holmdel, New Jersey, where the Bell Laboratories were developing under Project C-16 (summarized in Chapter 1 of this volume) a shielded-U Adcock for fixed-station service. Here polarization error measurements were made with steeply downcoming waves. Some description of the Holmdel equipment will be found in Chapter 1. The test transmitter was hoisted to the top of a 50-ft tower at Holmdel and hung approximately in line with the east-west Adcock pair described in the C-16 summary. Measurements were taken at six frequencies from 3.46 to 17.30 mc, with the test transmitter hung from the tower at 1.5° intervals from an elevation of 1.5° to 13.75° and when suspended by the balloon to elevations corresponding to 50° or 60°.

At each frequency and elevation, output of both Adcock antenna pairs was recorded both with the transmitter dipole vertical and with it horizontal. Unexplained minima of unwanted pickup for a transmitter elevation of about 5° were observed at all frequencies and were very pronounced at the higher ones; no corresponding horizontal field minima were observable.

Vertical to horizontal field-strength ratios at the center of the Adcock system, both for the Project C-17 tests and those at Holmdel, were computed using a number of terms of the series-expansion solution of Maxwell's equations given by Burrows.[4] The results indicated a tremendous enhancement of vertical field under the short-range transmission conditions used in the tests. Therefore, standard-wave errors for distant signals as determined from the above computed test-signal fields were much greater than such errors as commonly determined directly from measured ratios of wanted output for vertically polarized signal to unwanted output for horizontally polarized signal.

The directly measured results indicated that the Holmdel (C-16) Adcock was markedly less subject to polarization errors than the elevated-counterpoise Adcock of Project C-17, and was of the general quality (2° to 10° apparent standard-wave error in the range 17.5 to 3.5 mc) which other recent work had shown to be typical of good direction finders. Similar results for the C-17 Adcock with the better counterpoises ran from 7° to 15° in the frequency range 7.5 to 17.5 mc.

Extreme enhancement of local vertical fields is a matter of such tremendous importance to d-f testing, since it would completely invalidate almost all previous work, that it was studied further as reported under Project C-57. Approximate but seemingly sound application of general field theory to results of the Holmdel tests indicated improbably large standard-wave errors on distant signals.

8.8.2 Conclusions

The final report[2] on Project C-38 includes further general discussion of d-f design principles and testing methods, which leads to a number of conclusions.

Optimistic beliefs resulting from earlier work on the freedom of Adcock systems from polarization errors were not borne out by this or other recent work. In agreement with recent results of others on H Adcocks, it was concluded that shielded-U Adcocks are subject to a first-order error source of nature still unknown. In particular, the elevated counterpoise shielded-U system of Project C-17 did not compare as unfavorably with other systems as was at first supposed, so conclusions from its study are given in the form of concrete proposals for counterpoise design.

Since no direction finder can work well with all types of waves received, a "directive di-

versity" system was proposed in which some one of a group of two or three spaced-antenna direction finders, at the same location but each using a different type of antenna, will respond accurately to any coherent signal received. Use of devices to warn against vertically downcoming signals was suggested.

Knowledge of the means whereby polarization errors arise was not sufficient at the time the work was done to permit either sound design of direction finders or safe extrapolation from errors measured under usual test conditions to determine performance under widely different operating conditions. Inevitable presence of the ground improves performance of wave collectors at certain heights and injures their performance at other heights; whether good or bad, the effect is stronger the more conducting the ground. In general it was concluded that improvement of the direction finder itself was more to be desired than an equal improvement by choosing a site on better ground.

PROJECT C-57[a]

The great importance of having a truly reliable method of determining polarization errors, because of their probably larger magnitude in practical equipment, made it desirable to continue the work undertaken in the previous projects and to examine the experimental methods and the theoretical calculations of wave-field components used in testing under those projects.

The startling nature of the theoretical results obtained under Project C-38,[3] which appeared to invalidate practically all prior d-f measurements, indicated the desirability of a more thorough study. Thorough examination of the exact series-expansion solution of Maxwell's equations given by Burrows,[4] from which the approximations used in Project C-38 were obtained, showed it to be unsuitable for computation under just the conditions for which extreme enhancement of local vertical fields had been computed and reported under that project.[3]

A new approximate solution of Maxwell's equations, suitable for computing under the conditions of direction-finder testing, was developed from the exact solution in integral form given by van der Pol.[5] Comparisons with unpublished work of K. A. Norton showed this solution to be fundamentally the same as the one recently reported by him. Both solutions are valid under the short-range, high-angle conditions of d-f testing and both assume high ground conductivity. The new solution, in the relatively simple form given it by Norton, was used to re-evaluate the Holmdel results of Project C-38 and to analyze new experimental results obtained under Project C-57. In each case, the vertical electric field component produced near a horizontal rod antenna by curvature of the wave fronts was computed, as well as the horizontal electric field of the horizontal rod antenna and the vertical electric field of a vertical rod antenna.

Application of these results to the Holmdel data of Project C-38 showed clearly that no reliable measurement of polarization error of the Holmdel Adcock had been obtained. Spurious vertically polarized signal due to wave-front curvature near the horizontal test source had obscured the unwanted horizontal field pickup of the Adcock. This field curvature was evidently also the cause of the apparent minimum of error found at Holmdel for waves arriving at 5° elevation. A few of the balloon observations appeared to exhibit real polarization errors and permitted a rough estimate of standard-wave error as varying from 9 to 6½° between 5 and 9 mc. Pickup ratio, where determined, is apparently very low; good operating accuracy results from placing the system right on the surface of good ground. Some data taken by Bell Telephone Laboratories at Holmdel with both rod- and loop-antenna sources of test signal showed the same curvature effects. Up to the time this work was concluded, no measurements made on the Holmdel Adcock had been good enough to give an accurate determination of its polarization errors.

Because a horizontal loop transmitter does not produce spurious vertical electric fields due to wave-front curvature, further tests were made on the C-17 elevated-counterpoise Adcock to compare such a source with the horizontal

[a] Contract No. OEMsr-838, Radio Corporation of America.

rod or electric dipole radiator previously used. The rod-shaped test oscillator built for Project C-17 was modified to work with either rod or loop antenna and comparable tests were made with both source types.

No difference was found between measurements made on the elevated-counterpoise Adcock with rod and loop transmitters. The field computations indicated that real polarization errors were measured and were so great as to obscure the considerable field curvature effects. Pickup ratios were very low, especially for signals arriving at high elevation angles. They were of the same order as those estimated for the essentially similar Holmdel system, but the aid given to overall accuracy by good ground at Holmdel was lacking for the elevated-counterpoise system as measured at Princeton. Standard-wave error at 7 mc and over rather poor ground was found to be 32°.

Special tests of the loop transmitter showed that field-curvature effects were not eliminated but were reduced at least five-fold by its use. The rod transmitter is inadequate for measuring standard wave errors below 20°, except for high elevation angles, while the special loop transmitter used in this project should measure reliably errors as small as 4°.

Introduction of damping resistors into the elevated-counterpoise structure failed to reduce errors but did show how size relations between conductors acted to equalize errors over a wide frequency band. No evidence could be found that voltages induced in the counterpoise by horizontally polarized signals were fed into the antennas by capacitance, but it was noted that a single vertical antenna mounted eccentrically above the counterpoise was markedly more responsive to horizontally polarized signals than was a centrally located vertical antenna.

Ordinary testing equipment and methods are clearly inadequate for the study of polarization errors of really good direction finders. At the close of this work, it appeared that no fully adequate test had yet been made of the downcoming wave performance of any very good direction finder. Vertical field enhancement near a local transmittter is not as extreme as the result of Project C-38 had indicated and is unimportant at high angles. It is quite important at the low elevations and short ranges used in much d-f testing.

5.4 PROJECT C-78[a]

Project C-78[a] was concerned with the measurement of errors of radio direction finders and served to correlate and evaluate knowledge of measuring techniques gained in the work of Projecte C-17, C-38, and C-57. The whole problem of such measurements was surveyed including the question of what to measure, how to measure it, of the range of measurement necessary or desirable, and of the characteristics required in the measuring equipment. Because of the great importance of the technical capabilities of radio direction finders to their user, methods of performance testing require careful specification. Overall tests designed to simulate actual operating conditions and to yield direct information as to accuracy of bearings are highly desirable.

Noise level and accuracy of equipment auxiliary to the d-f antenna system, like accuracy of reading at various steady signal levels, can and should be separately determined by normal laboratory methods. Conditions for determining reading errors on actual fluctuating signals cannot readily be specified. Errors due to good signals arriving by laterally distorted paths are not errors of the direction finder itself, while signals arriving from elevations above about 60° should not be used for direction finding. Testing methods must avoid conditions which cannot be specified clearly or should properly be excluded from measurement.

Errors due to interference among signal components arriving over multiple paths are of great importance, as are errors caused by electrical inhomogeneities in the immediate surroundings of a dire[illegible] finder. Conditions for measurement of th[illegible] errors could not yet be specified at the co[illegible]usion of this project. Test methods should avoid producing such errors, yet be adaptable to their controlled production when the art permits specification of appropriate tests conditions.

Failure of actual d-f wave collectors to re-

[a] Contract No. OEMsr-838, Radio Corporation of America.

semble exactly their ideal prototypes, even when well located and receiving a steady single-component signal, was an important source of error at the time this work was done. Conditions for measurement of such errors could already be specified and they could and should have been measured reliably, but this had practically never been done. These errors are of two types often called "calibration" and "polarization" errors (night effects), and it was to their measurement that Project C-78 was directed.

Actual distant signals have very complex properties which are usually incompletely known and are therefore poorly suited for performance testing, though limited data can be obtained statistically from large numbers of distant-signal observations. Reliable and complete testing requires a fully controllable local source of test signals, arrenged to simulate the properties of certain typical distant signals. A few actual distant-signal observations are desirable to check the validity of the local-source test method.

Measurements made with local sources under simplified limiting conditions may not always be reliable guides to performance under all operating conditions. Even when special detailed knowledge of a particular direction finder permits generel conclusions to be drawn from simplified measurements, the rigorous theoretical work required may be less convenient than more complete direct measurements. The pronounced effect of the inevitable presence of the earth near every direction finder cannot always be treated as separate from its intrinsic performance; in some cases only overall performance of ground and direction finder together is significant.

Carefully interpreted measurements from a lower limiting frequency between 6 and 10 mc to an upper limit between 18 and 30 mc can indicate performance of similar direction finders over the entire h-f band from 1.5 to 30 mc. Model measurements at very high frequencies avoid some difficulties of full-scale testing but require development of suitable models and may be misleading because the model does not accurately simulate obscure imperfections important in the original.

Variation of error with signal-arrival azimuth on favorably polarized signals and variation of error with signal-arrival elevation on unfavorably polarized signals must both be determined at several frequencies. The results can only be fully shown as graphs, but effort should continue to express a maximum of information by a few figures of merit. Such measurements should be carried out over very uniform highly conducting ground to determine ultimate performance capability and over uniform poorly conducting ground as well to determine possible impairment of performance.

The primary instrument used in d-f testing is a signal field and test methods must be planned on the basis of accurate knowledge of its characteristics. Approximate expressions defining this field, as developed by Norton, show its properties to be quite complex, especially near the signal source. This complexity is caused mainly by the presence of the earth's surface.

Signals from a local source differ from those from a distant source in two ways. The local signal shows different rates of attenuation with distance for components plane-polarized respectively parallel and perpendicular to the vertical plane of wave travel, while the distant signal shows no such difference. The local signal, spreading from a small source, has curved wave fronts which cause somewhat different fields to appear at laterally separated parts of the direction finder, while the distant signal has plane-wave fronts. Each of these differences can seriously confuse d-f error measurements. Both can be avoided only if all measurements are made at transmitter-receiver distances of at least several tens of wavelengths.

The main method of measurement used in recent work involves two observations of the output of the d-f wave-collector system under test, one in a field of the polarization to which the collector elements were designed to respond and the other in a field to which no response was intended. The ratio of these outputs, for equally strong incoming signals, gives the maximum polarization error to be expected. Separate observation with signals of limiting polarization avoids phaeing difficulties of earlier work where both signals were present at once, but requires test signal sources of extreme polarization purity.

Projects C-17, C-38, C-57, and C-78 used a signal source designed to achieve pure polarization by being entirely self-contained and conforming in outline to the intended antenna, a short rod (electric dipole) or small loop (magnetic dipole). The frequency was stabilized by crystal control of a master oscillator driving a balanced power amplifier coupled to the highly reactive antenna by an autotransformer. This source was light in weight for convenience in elevated operation and its power output was maximized by use of efficient miniature power tubes and miniature batteries under very heavy load. Figure 2 shows a close-up of the interior of the 4-in. by 2-ft cylindrical test source, with inserts showing its incorporation into rod and loop radiators.

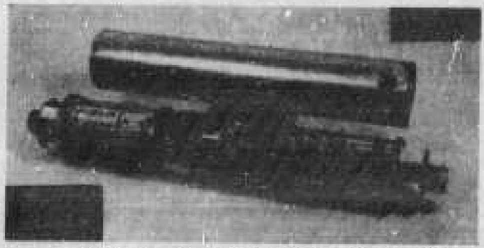


FIGURE 2. Interior arrangement of test source employed in Projects C-17, C-38, C-57, and C-78.

The usual method of determining purity of transmitter polarization, by observing the output of a receiving antenna of supposedly pure polarization with the transmitter in various orientations, was shown by a complete analysis to be generally incapable of giving the desired result. Tests of this type, using an accurately vertical rod receiving antenna centered over an accurately horizontal circular elevated counterpoise, were made on signals from the above source and indicated rather disappointing performance. Unwanted emission, while not definitely determinable by this method, seemed to be at least one per cent of wanted emission in field strength. This purity is adequate for tests in which a direction-finder null may be used to discriminate against unwanted emission but quite inadequate for more exacting tests. An appreciable electric-dipole moment was exhibited by the magnetic-dipole source, probably because of the breaks in the loop required to connect the generator. Elimination of unwanted electric field components due to wave-front curvature, attractive in principle, was thus found very difficult in practice.

The output-ratio method of d-f testing is indirect and slow, beside requiring inconvenient manipulation and sometimes needing an unattainably pure source. Some other means of avoiding error reduction by chance favorable phasing of various field components would avoid these difficulties. An improved method of d-f error measurement based on a novel test signal source was proposed in the final report on Project C-78.[a] The proposed signal source would use an antenna unit consisting of two distinct radiators of different polarization, preferably a vertical rod and horizontal loop. These would both be fed from a common r-f

generator, with constant relative amplitude and continuously varying relative phase. Polarization error would be observed as a swinging bearing and measured by amplitude of swing. Error measurement would thus be direct, rapid, and experimentally convenient, and no critical control of orientation of highly elevated equipment would be necessary. Slight polarization impurity would cause only small inaccuracy of error determination, instead of seriously obscuring the significance of the results; careful design of the test source would nevertheless be required to maintain fairly good purity.

At the large distances so necessary to insure freedom from confusing local-field effects, full freedom of control of position of the test source is only possible by supporting the source from an aircraft. Airplanes are not convenient for such work but a nonmetallic dirigible airship would be very valuable. Captive balloons are inferior to dirigibles but perhaps more practical; they should be of good aerodynamic form, be lightly loaded and carry a source which does not require adjustment of orientation in flight. A captive balloon has been found quite useful even though all three of these conditions for satisfactory operation were violated. Tall towers or poles provide very convenient support but their range of usefulness is necessarily limited.

Sites for testing d-f performance must be much more critically chosen than even d-f operating sites. They must be clear, flat, and electrically homogeneous over a radius of several tens of wavelengths at the lowest frequency to be used and to sufficient depth to attenuate the transmitted wave by ten times. Wastelands are fortunately very suitable, salt marshes as sites of high conductivity and deserts as sites of low conductivity.

By use of a source of the type proposed supported from an aircraft over well chosen sites and working at adequate distances, d-f performance should be assessable with an ease, completeness, and reliability not approached in any tests hitherto made. Tests by these methods can be extended to include effects of multipath wave interference and of inhomogeneous sites if the art advances sufficiently to permit appropriate test conditions to be specified.

8.8 PROJECT C-58[a]

8.8.1 Causes of "Swinging" Bearings

The original development of the Adcock antenna system was for the purpose of rendering an associated direction finder insensitive to that component of a radio wave whose electric field is polarized perpendicular to the plane of incidence (horizontally polarized). Theoretical computations, as well as tests with a controlled local target transmitter, indicated that the Adcock antennas developed under Project C-34[b] were capable of discriminating to a very high degree between the desired vertically polarized and undesired horizontally polarized waves of a radio signal. Nevertheless, tests on sky-wave transmissions revealed swinging bearings on the cathode-ray indicator of the direction finder, typical of so-called polarization error. Both the magnitude of the bearing oscillation and the percentage of time that the cathode-ray indication departed from the correct azimuth made it appear likely either that theoretical computations of wave discrimination for these antennas were grossly in error, or that the downcoming sky waves were not polarized at random according to the generally accepted hypothesis of ionosphere reflections, if it were assumed that the swinging of the bearing was due to polarization error.

Because the values of polarization discrimination by Adcock antennas were more or less substantiated by tests with local target transmitters, while the distribution of polarization of downcoming sky waves remained unproven by any physical tests whose results could be directly associated with the apparent polarization of Adcock direction finders, the desirability of tests on the polarization of downcoming sky waves was clearly indicated. A part of Project C-58[a] concerns the study of polarization of radio waves between 5 and 20 mc.

In agreement with this contract, there was built and installed at Great River, New York, an equipment since called the polariscope, a description of which follows. This equipment permits the ratio between the vertical electric

[a] Contract No. OEMsr-745, Federal Telephone and Radio Corporation.

and horizontal electric components of a radio wave to be seen at a certain point while radio bearings are observed on a cathode-ray indicator also to be described below.

The Bureau of Standards, aware of these facts, asked the contractor to observe bearings and polarizations of the Bureau's station WWV, Beltsville, Maryland, transmitting successively with two different types of antennas, radiating at certain times vertical and at other times horizontal polarization. This transmitter is sufficiently distant from Great River that no ground wave is present.

8.8.2 Description of Polariscope

The circuits and antenna design of the direction finder are approximately the same as described in Chapter 9 dealing with the SCR-502 (Project C-34).[10]

A block diagram of the polariscope and d-f equipment is given in Figure 3. The Type A indicator is identical with the d-f indicator used in SCR-502. Figure 4 is a photograph of the antennas used with the polariscope, and these antennas consist of two crossed dipoles 12 ft in length mounted on a revolving boom 20 ft long. This entire boom with its central column can be revolved from within the control room so that it may face the direction from which the signal arrives. This assembly is mounted on a small wooden tower with its central axis about 20 ft above the surface of the earth.

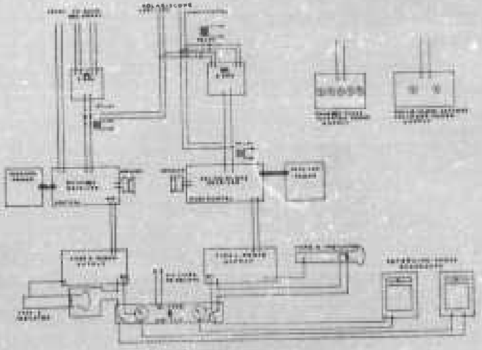


FIGURE 3. Block diagram of polariscope.

Each antenna, both horizontal and vertical, has its own balanced cathode-follower coupling unit, which in turn feeds into a balanced dual coaxial cable leading into the central control room about 50 ft away. The vertical antenna is connected to the vertical stator of the goniometer, and the horizontal antenna to the horizontal stator of the goniometer. In this manner,

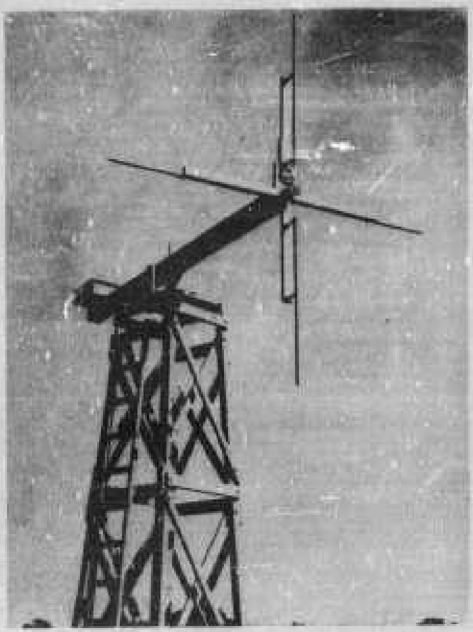


FIGURE [illegible]. Antenna used with polariscope studies.

the pattern developed on the face of the indicator unit indicates the amplitude and phase relationship between the vertical and horizontal components of the electric field of the received wave.

In addition to the visual indication on the cathode-ray oscilloscope indicator, automatic recordings of the amplitudes of the two components of the wave are made by two Esterline-Angus recorders, with a paper speed of 3 in. per minute, set up in conjunction with two separate receiving channels fed by the two dipoles of the polariscope.

The program for the reception of WWV was to observe the bearings and the polarization for 5 minutes; then to align the two receivers alike and record the intensities of the two components for 5 minutes. After this the bearings and polarization were visually observed for 5 minutes again. The whole procedure was repeated every 15 minutes for each type of transmission. The last half of each hour was used for a standby period, during which time the two receivers were recalibrated for identical sensitivities with the next frequency to be tested.

Analysis of WWV Observations

Three days' operation of this equipment resulted in indications and records of bearing errors with wave polarization, the analysis of which is as follows:

1. The horizontal vector of downcoming waves from both the vertically polarized transmitter and the horizontally [illegible] transmitter was found to have random polarization. This was in accord with the generally accepted theories on polarization of sky waves at these frequencies.

2. During periods when the sky-wave polarization was horizontal or a very few degrees from horizontal, bearing indication was in error or indeterminate. This was the expected polarization error.

3. Frequently, even when the sky wave was vertically or nearly vertically polarized, there were wild oscillations of the bearing indicator and deterioration of the null.

The above results indicated that the oscillations of the d-f bearing were not due solely to polarization error as had been assumed. The oscillation of the bearing during periods when the wave was approximately vertically polarized must be due to some other phenomenon. The following hypothesis and tests were an outgrowth of the analysis of the above polarimeter investigations.

Wave Interference Errors. The hypothesis which assumed swinging bearings to be due to strong horizontally polarized components in the downcoming sky wave, in general also assumed that the wave was reflected from a single point in the ionosphere. Were this the case, then swinging bearings would feasibly be due only to horizontal polarized sky waves.

But consider the result of combined waves from two or more points in the ionosphere. If these reflection points differ even by a degree or less, the combined electric vector at the d-f antenna will be a function of the instantaneous phase difference between the several combined waves.

An analysis of more than two rays becomes extremely involved, therefore the combination of only two waves will be discussed here. The result of the combination of several rays which, in practice, frequently arrive at the direction finder from a single transmitter, will be a still greater variation in the bearing indications.

Figure 5 is the special example of two rays whose azimuths differ (for the sake of clarity) by a much larger angle than that usually experienced in practice. In practice, relative magnitudes of the separate waves vary as do their instantaneous phases. This is due to the fact that the separate rays apparently arrive from regions in the ionosphere which differ in height and density of ionization and where ionization conditions are not necessarily stable.

A simple example of the mechanism whereby two vertically polarized waves from slightly different azimuths can result in a large d-f error is as follows:

First consider ray C in Figure 5 to arrive at point O with its instantaneous error vector directed upwards and ray D to arrive at the same point with its vector similarly directed upwards. As long as these vectors remain in such phase, a direction finder located at point O will provide an indication between C and D. (Since in practice C and D usually differ by a

very small angle, this may be called the correct bearing indication.)

Now consider a later time when the ray from C, while still vertically polarized, has altered its phase with respect to the ray from D by 180°. Then the desired vertical components tend to cancel while the component of the vector which bisects the angle between C and D tends to be additive. In such a case, if the amplitude of the waves from C and D were identical, the signal would be very weak and the d-f bearing indication would be 90° in error. With unequal amplitudes, the error is less than 90°.

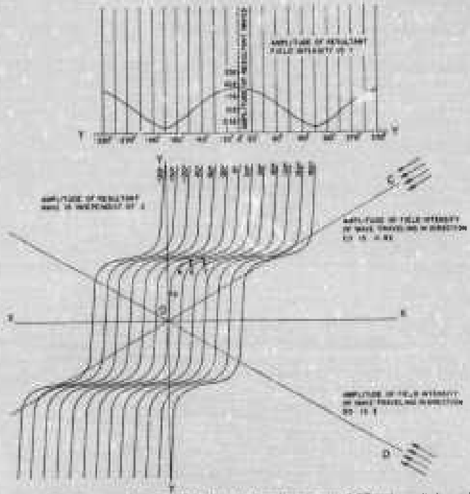


FIGURE 5. Equiphase curves for resultant of two vertically polarized waves of different magnitudes moving in different directions.

In a typical situation, the combination of two rays will occur with varying phase and relative amplitudes, thus providing an oscillating indication of bearing. The combination

of more than two rays will, of course, increase the complexity of the resultant vector to which the direction finder responds.

Because in the crossed Adcock direction finder there is in general a small azimuth error which varies with the vertical angle of incidence (the octantal error), two rays arriving from the same azimuth direction, but from different layers of the ionosphere will also result in an interference error whose behavior is similar to that due to two rays arriving from slightly different azimuths.

Test of Wave Interference Errors. To examine the error resulting from the interference of two waves whose azimuth and relative phase may be controlled, the direction finder employed in the previously described polarization test was used in the reception of signals from two local target transmitters.

These two transmitters were located about 1,000 ft away from the Adcock antennas of the direction finder, and were spaced so that the azimuths of arrival differed by about 2°. One of the transmitters was set on a given frequency (about 5 mc), while the other transmitter frequency was varied by hand to be as close as possible to the frequency of the fixed transmitter.

As the frequencies of the two transmitters fell within the band width of the direction finder, a confused and rapidly changing fluctuation was observed which in general pointed toward the two transmitters.

When the two transmitter frequencies were brought as close together as was possible (for brief periods two frequencies were apparently within 1 or 2 cycles), the d-f indication was that of a slowly oscillating bearing whose quality was highest when it pointed in the direction of the two transmitters and which deteriorated to almost no indication when it approached a bearing 90° from the line bisecting the angle between the two transmitters.

It was suggested that a further experiment to investigate precisely the errors due to any given phase difference and amplitude between the two incoming rays could be performed by feeding two displaced transmitting antennas from a single transmitter with a phase and amplitude adjustment in the line to one of the antennas. But, because the test with the two separate transmitters satisfactorily proved that two vertically polarized waves arriving from slightly different azimuths can cause bearing oscillations resembling polarization errors, it was decided that, in view of the fact that a combination of only two rays was artificial, further tests of this type should not be pursued at this time.

Conclusions

The experience with the polariscope and the later tests which simulated two sky waves have indicated that the d-f bearing oscillations and deterioration nulls are not due entirely to polarization error.

In America it has been assumed that efforts to disseminate horizontally polarized waves would ultimately result in a direction finder whose bearings would be steady and precise beyond those which had been designed in the past.

The facts that the prevalent multiple-ray transmission of radio signals gives rise to interference errors in Adcock direction finders would indicate that too great effort in reduction of polarization error are not warranted.

At present, it appears that the interference error cannot be reduced by any d-f system in which the antennas cover such a limited area as do the Adcock antennas. The Musa (multiple unit steerable antenna developed by Bell Telephone Laboratories), does reduce interference errors. The Musa system, however, is necessarily a very large installation which can be used for direction finding over a very small azimuth only.

8.6 PROJECT 13.1-84

Under Project 13.1-84[f] a very great deal of work was done to determine the essential characteristics of the ground under d-f installations to make the apparatus as useful and as free from errors as possible. Part of the project was to develop, if possible, simple equipment which a relatively untrained person could take to a site selected for a d-f installation, perform a simple and not too critical experiment, and by means as simple as reading a meter, perhaps

[f] Contract No. OEMsr-1026, Federal Telephone and Radio Corporation.

like a tube-testing meter, determine whether the site was suitable or not. All previous work under Division 13 projects had indicated the extreme importance of locating d-f apparatus on sites with good ground characteristics.

The final report[11] on Project 13.1-84 shows that the phenomena involved are too complex for a single instrument of simple type to be constructed for the job to be done.

An extensive bibliography is contained in the final report[11] and this report should be consulted by anyone seriously engaged in site investigations. The bibliography has references to characteristics of the soil as of interest to a chemist, or from the standpoint of electrochemistry or physical chemistry. The report itself contains much historical material dealing with our knowledge of the conductivity of the soil, its dielectric behavior, its d-c and h-f resistance, and of methods explored for determining these factors.

8.6.1 Methods of Measurement Studied

A comparison of resistivities as measured by direct currents and by alternating currents indicates that electrode effects cause the d-c resistivities measured at high radio frequencies before a dispersion occurs.

Nevertheless the final report indicates that d-c measurements may be a very practical method for determining resistivity of soil samples.

R-F Measurements

Considering the soil as an imperfect dielectric, a whole series of experiments was performed to determine the relative qualities of soils as dielectrics. By the use of a Q-meter, in which the soil is inserted between the plates of a capacitor and the overall loss factor of soil plus capacitor determined, the conclusion was reached that Q-meter measurements were not practicable over a wide frequency range. The method is not suitable for measuring resistance of sample soils greater than 50,000 ohms, the reliability of the method increasing as the resistance decreases.

The conclusion was reached that the Q-meter method could be used at spot frequencies for soil samples in Lucite containers.

Since it is well known that the inductance of a coil at radio frequencies depends upon the core material, the Q-meter can be used to measure soil characteristics by placing the sample inside a coil whose inductance, without the soil, is known.

It was found that measurements by this method did not give quantitative results but with care could be made to show relative quality. The inductance method is more sensitive and more easily applied than the capacitance method.

Bridge Methods

The most accurate way of measuring the impedance of a soil sample at radio frequencies is to use a suitable bridge circuit. This method requires more skill, is more tedious, and the range of values measurable is less than in some other methods. In addition an auxiliary generator and detector are required.

Because of variations of the weather and the averaging of soil constants in wave propagation, it seems unnecessary to measure the conductivity with an accuracy greater than 50 per cent of the mean value of several measurements. Furthermore, measurement of conductivity alone seems to be all that is necessary to determine the characteristics of ground material of a possible site for a d-f installation. Thus the bridge method, while yielding a high degree of precision, is too complex for the job to be done.

Methods Using Antennas and Transmission Lines

One of the most effective means of determining the effect of a site upon a d-f installation is to set up a portable direction finder of known properties and to observe how its operation is affected by the site.

Thus an antenna and its characteristics are a function of the ground upon which it is erected. Its input impedance, the ground losses, and directional patterns being functions of the ground. Properties of transmission lines which are most susceptible to investigation are attenuation constant and velocity of propagation. Preliminary measurements indicated that such studies could be quite effective but the decision was reached that more work would be neces-

sary to find out if the methods would be practicable. Furthermore, such experiments could be performed only with engineering supervision, such was the state of the art of measuring equipmant of the kind required for antenna or transmission-line measurements.

Similarly, measurements of field strength at different frequencies were abandoned when it was found that results were inconclusive. Reflection coefficients and wave tilt were studied as a function of ground constants. Limited experience indicates that the method is applicable to measurements of soil constants under conditions of (1) no obstructions between transmitter and receiver, (2) no reradiators near enough to cause trouble, (3) elevation of the target transmitter above ground or else use of rather large power input. These limitations were discouraging from the standpoint of portability and the necessity of determining ground characteristics under varied conditions.

Audio-Frequency Methods

Mapping a site by plotting equipotential lines between ground electrodes at audio frequency required less time than mapping by plotting equifield lines about a transmitting antenna at radio frequencies. The use of audio frequencies and ground rods eliminates pronounced disturbences caused by above-ground reradiators observed in plotting equifield r-f field lines. The method employed in Project 13.1-84 is described in the final report [11] and is applicable to the picking of a site for a d-f station. The method can be used, also, for locating large bodies of metal under the surface of the ground and this is discussed in the final report, together with the use of r-f devices such as mine detectors for locating small metallic bodies. Circuits for such devices as constructed under the project will be found in the final report.

Wenner-Gish-Rooney Method

In this manner the following process is carried out: four copper-coated rods ⅓ in. in diameter and 1 ft long are used as electrodes. They are equally spaced along a straight line and voltmeter readings are taken for several values of spacing between 1 and 35 ft. A battery-operated vibrator delivering approximately a square-wave alternating current of 110 volts is connected through a milliammeter to the outer electrodes and a battery-operated vacuum-tube voltmeter is connected between the inner electrodes. The current flowing through the outer electrodes and the voltege between the inner electrodes are read for each of the chosen spacings. At close spacings the electrodes are driven into the ground only an inch or so, at greater spacings the electrodes go into the ground to depths of up to 1 ft.

In this manner a plot of an area showing effective resistance as a function of depth can be obtained. The method is easy to apply, is sufficiently accurate in indicating the resistivity of the top surface of the ground and is the best method of obtaining in a qualitetive manner the resistance as a function of depth.

The final report ends with some date on the ground requirements for direction finding in various frequency bands, indicating in a particular case that 50 tons of coal dust screenings, either soft or hard coal, should be put down to a depth of 1 in. under installed ground mats and to a radius of 10 ft beyond the guy wire anchors. The ground mats and the coal dust layer are covered to a thickness of 3 in. and this is tamped down tightly.

Treatment of this sort produced a ground which contributed little trouble to the d-f station involved.

8.7 PROJECT C-19[12]

The loop direction finder has been found lacking as a dependable instrument for navigational and other purposes because of inaccuracies under certain operating conditions. It is often impossible to get a bearing at all or the azimuth of the observed bearing may be greatly in error or may vary from moment to moment. These errors have been under continual study since the loop direction finder came under practical use during World War I and the basic causes for the different types of errors are now well understood. Most of the errors have proven capable of elimination, but a notable exception has been polarization error which includes the so-called night effect and airplane

effect. Project C-19[e] was to study a particular and new means for attacking this type of error. Project 13-122[13] studied this compensation means critically and reported on difficulties with it.

5.7.1 Normal Loop Operation

The ideal case for loop operation is a vertically polarized wave (in which the electric vector is always in the vertical plane through the direction of travel) proceeding along the surface of the earth and following a great-circle path between the transmitter and the receiver. In this situation the loop has the well-known "figure eight" directional response, a minimum or null being obtained when the plane of the loop is at right angles to the direction of arrival of the signal.

Under these conditions and if the loop system itself has no "instrumental" errors, the bearing of the distant station can be ascertained with considerable accuracy.

5.7.2 Wave Errors

If, however, there are abnormalities in the wave itself, a loop which will operate perfectly on normal waves will show errors in bearing, hazy bearings or no bearing at all.

The several wave errors are as follows:

1. Coastal refraction is the phenomenon resulting when the received signal travels obliquely across a boundary between two soil types, notably ocean-to-shore transition at a d-f site located some distance from the coast line. The wave is actually refracted and appears to come from an incorrect direction.

2. Lateral deviation is a phenomenon in which the wave does not travel a great-circle course but deviates as much as 10° from this course.

3. Scattered signals is another form of wave error in which the signals seem to arrive from severel directions, apparently from scatter sources in the ionosphere or on the earth's surface which appear to reradiate some of the original energy.

[e] Contract No. NDCrc-159, Stanford University.

5.7.3 Polarization Errors

Far more common than the anomalies of scattered signals and lateral deviation are the errors due to irregular polarization of the received wave. The symptoms are of several types as follows: (1) Bearing sharply defined and stable but apparent azimuth incorrect; (2) bearing sharply defined but shifting in direction over a period of time, often quite rapidly; (3) blurred, indefinite null point, although a minimum of correct bearing may be detected.

Polarication errors occur when the received wave is not a simple, vertically polarized wave but contains a horizontal component as well. This horizontal component arises from the rotation of the plane of polarication of the sky wave in its reflection from the ionosphere. When such a wave arrives at a d-f station, the operator turns the loop to get a null indication but is able to do so only when he has oriented the loop in such a manner that loop voltages due to the vertical component (vertical loop conductors) and due to the horizontal component (horizontal loop conductors) are equal and opposite. This is not the loop position which gives a null on the vertical component only because the angle of arrival is such that there is a phase difference between the two components. Therefore, the operator gets a wrong bearing.

5.7.4 Attacking the Problem

Two possible modes of attacking this problem present themselves. One possibility is to design a collector with only vertical elements. This leads to the Adcock antenna which is quite useful for many applications. Its great disadvantage is the fact that its pickup, unless the structure is quite large, is small whereas the loop can have many turns with correspondingly greater sensitivity.

The mode of attack pursued under Project C-19 was to accept the situation of having troublesome errors due to the horizontal pickup but to compensate the unwanted voltage by another horizontal voltage secured from an additional antenna mounted with the loop and rotating with it. This forms the so-called compensated loop which has been discussed in the literature and on which patents have been granted.[14]

In the system proposed, the voltage induced in the auxiliary antenna would be expected to behave in the same manner as the voltage induced in the loop by the horizontal polarization. Then this voltage would be coupled into the loop in such a manner as to provide neutralization for the unwanted voltage.

Two basic problems are to be solved. First, what must be the network characteristics used for coupling the neutralizing voltage into the loop and, second, to what extent does the neutralization become incomplete if one of the operating variables change.

The bulk of the final report is devoted to a study of these basic problems including the effect of a wave reflected from the earth's surface, the effects of vertical and horizontal polarization or a combination of the two, errors in the uncompensated loop and in the compensated arrangement, calculations on typical situations, the effect of wavelength on compensation, variation of height of antenna above ground, and height of auxiliary antenna with respect to the loop. There is considerable material relating to the measurement of ground reflection coefficients, voltage ratios and phase angles, field strength, etc.

5.7.5 Results Obtained

As a result of the theoretical analysis and extensive field tests, it is concluded that the system would work for short as well as for long waves, calculations being given for a range of from 1- to 1,000-meters wavelength, that it will function for any type of soil condition, that it will not work on airplanes where extreme changes in soil type would occur over which the plane flies or where large variations in height above ground would occur. The system works best at fixed heights which are small ($\lambda/10$ or less) compared to the wavelength.

On actual demonstration of experimental equipment and a Sperry Mk-I automatic direction finder good compensation was secured.

An extensive bibliography is included in the final report.[12]

5.7.6 Compensated-Loop Direction Finder

The Signal Corps of the U. S. Army in January 1944, requested NDRC to perform research on a loop antenna satisfactory for direction finding on transmitters up to 30 miles away in the h-f band which would be as satisfactory at nighttime as during the day.

Under the continuing Project C-58,[a] some investigations were made on compensated-loop direction finders.

In the past considerable work has been performed in attempts to compensate loop antennas against response to horizontally polarized waves. In almost all such investigations the general problem of downcoming sky waves at any angle has been attacked. The failure to design a satisfactory compensated-loop direction finder may have been due to the too general nature of the problem.

A loop direction finder compensated against night errors for transmissions of not more than 30 miles would require that compensation be against vertically or nearly vertically downcoming waves only. It might be expected that this special problem could be solved more easily than the general loop compensation which had as yet no satisfactory practical solution.

Although the final form of compensated loop might for reasons of portability be a single rotatable loop antenna with the necessary attachments for response to horizontally polarized downcoming waves, it was decided that for reasons of convenience during the experiments a fixed crossed loop be employed.

Directly below the crossed loops were installed crossed horizontal dipoles. An injection-loop transmitter was located 30 ft directly above this collector, to generate the vertically downcoming wave.

Both the loop antennas and the horizontal dipole antennas fed cathode-follower coupling units. In the dipole coupling units both amplitude and phase were adjustable.

Experiments were made at night on a transmitter located 25 miles away. The loop transmitter located above the collector assembly was tuned to the frequency of the distant transmitter and then the compensating dipole antenna coupling units were adjusted to minimize the signal. A reduction of about 10 db was easily accomplished. The distant transmitter was then turned on, and it was found that on the cathode-ray indicator the swinging of the bear-

ing was reduced from four to eight times as compared with the uncompensated loop.

It was found that the improvement was best when the injection loop transmitted at precisely the same frequency as the distant transmitter. Also, an adjustment made when the ground under the loops was dry became worthless when a brief rain altered the conductivity of the ground. On the whole, the compensation was not complete, and the apparent requirement for adjusting the coupling units by means of an injection signal exactly the same frequency as the distant signal was a serious limitation. During the experiment it was also found that a slight frequency shift by the injection transmitter required a large adjustment in the coupling unit controls.

In the May 1945 issue of *Proceedings of the I.R.E.*[15] an article by J. N. Pettit and A. W. Terman on compensated-loop direction finders concluded with some encouraging remarks on the possibility of compensating a loop antenna by means of a horizontal dipole. Because this conclusion apparently differs from that reached in the report of Project C-58 on compensated loops, the NDRC requested a comparison and discussion of the two reports to resolve the apparent contradictions.

8.8 PROJECT 13-122[19]

8.8.1 Discussion of Project C-19

Under Project 13-122,[h] a final report was prepared which discusses the work accomplished under Project C-19.[18] The gist of this discussion follows.

The compensated loop was studied under Project C-58 and the report on that project states that results were rather discouraging. The important item to be determined is whether the findings of Project C-19 were corroborated by the work in Project C-58 or whether there is some basic difference between the results. It is concluded that, basically, there is no theoretical disagreement. However, it is shown that the coupling networks should, if possible, include means for resolving the differences in the internal impedances of the loop and dipole antennas.

The report of Project C-19 on the investigation of compensation in direction finders is a mathematical investigation to determine the phase angle and the amplitude ratio between the voltage induced in a loop antenna by a downcoming horizontally polarized wave, and the voltage induced by the same wave in a horizontal dipole mounted at the center of the loop. It was shown mathematically that in the presence of grounds of medium conductivity, or better, this amplitude ratio and phase difference remain nearly constant for varying angles of incidence, and for varying frequency. For instance, with wet soil, between the wavelengths of 1 and 20 meters, the amplitude ratio varies from 6.4 to 6.8 and the phase shift varies from 9.9° to 8.5°. These are calculations of the voltages induced into the antennas and do not take into account the internal impedances of the antennas. Assuming that the voltages, once they were introduced into the antennas are available, the report shows that the compensation requirements vary slowly with frequency; and that for various types of soil, except very dry soil, the ratio of the two voltages and the phase shift between them remain constant, provided that the antennas are mounted less than $\lambda/10$ above the ground.

Coupling Network

It was concluded that it was necessary to design a circuit which would give constant phase shift, constant amplitude ratio, and good stability with varying frequency. Such a circuit was designed under C-58. For convenience of indication a crossed-loop system with two horizontal dipoles was used. An instantaneous cathode-ray indicator for bearing indication was employed. There were direct low-impedance connections between the loops and the goniometer. Each dipole antenna was then coupled to the corresponding low-impedance connection through a set of two balanced cathode-follower coupling units. One cathode follower operated without phase shift and the other cathode follower in the set had its phase shifted by 90°, so that by combining the two and changing their

[h] Contract No. OEMsr-1490, Federal Telephone and Radio Corporation.

relative gains, the output phase could be shifted from 0° to 90°.

Although this coupling unit is of the type that was indicated by the conclusion of the C-19 report, it does not take into account the varying impedance of the dipole antenna with frequency and the varying impedance of the loop antenna with frequency. To employ the ratio of the two induced voltages, it would be necessary, if it is at all possible, to obtain these voltages, for combination, without any phase shift, or amplitude ratio shift, introduced by either internal impedances or external added impedances in the antenna units.

The final proof submitted in Project C-19 was a d-f test at one frequency and at one downcoming angle. The direction finder was adjusted for good results with the target transmitter and it is shown that the type of polarization transmitted by the target transmitter does not thereafter introduce any error. This test was repeated in Project C-53 as stated in their report of July 1943.[16] For the purpose a polarized transmitter was installed atop a 90-ft tower. However, in the report of September 28, 1943,[17] on the problem of making the adjustments with the target transmitter, it is noted that without the help of a target transmitter producing a downcoming wave at the frequency of the transmitter to be observed, the various adjustments of amplitude and phase cannot be carried out with certainty, and that the practical development of such a system for the Armed Forces did not look promising.

Comparison of Reports

The final report on C-58 contains no findings in conflict with the results of C-19. The report of May 28, 1943 (C-58)[18] states that the phase difference seems to remain constant, but a great deal of difficulty is encountered in checking the amplitude relationships, since they seem to vary. It is also stated in the report of July 1943,[16] that "the phase and amplitude relationships remain constant over long periods of time and the various states of polarization." This finding seems to agree very closely with the theoretical calculations made under Project C-19.

Since it was necessary to work for a practical solution from the mathematical conclusions, it was necessary to investigate the amplitude and phase relationships between the two voltages to be opposed as a function of: (1) polarization, (2) ground angle of the sky wave, (3) frequency, and (4) ground constants.

Once these relationships were proven to be constant, or very nearly so, it was necessary to devise some circuits which could be adjusted easily and with certainty. In the report of September 28, 1943[17] it is stated that a target transmitter is needed for making these adjustments. This seems to be a very reasonable assumption unless the ground conditions can be measured (which would be rather unreliable, since the ground might vary over very large areas), and the adjustments to be made then calculated from those measurements. This solution did not seem practical for a useful military direction finder.

Conclusions

The investigations under Project C-58 on compensated loops revealed a problem not mentioned in the Pettit and Terman report. That is, the varying impedances of the antennas with varying frequency and ground conditions effectively prevent the use of the voltages induced in infinite-impedance antennas to compensate against horizontally polarized waves. The voltages discussed in the former report must be assumed to exist in infinite-impedance antennas, but such antennas are not available in practice. Since the loop antenna's principal value is that it may be tuned, and when tuned its impedance is a critical function of frequency, the conditions of infinite impedance for antennas in a compensated-loop system are not practicable.

Chapter 6

CORRELATION OF D-F ERRORS WITH IONOSPHERE MEASUREMENTS

PRIOR TO THE WAR no coordinated study of ionosphere transmission and d-f errors at high radio frequencies had ever been attempted. Such a study was desirable from the standpoint of determining the causes of deviations from great-circle transmission paths and to establish criteria for the presence and extent of d-f errors caused by the ionosphere.

At a series of conferences called by Division 13, NDRC, and beginning in late January 1941, plans were made for systematic observations of ionosphere characteristics and d-f errors in the range 2 to 30 mc in which waves are reflected from the ionosphere. The general plan was to have simultaneous ionosphere and d-f observations at a number of points on this continent. As a result projects were set up in Division 13 to implement this coordinated study. Numerous ionosphere laboratories furnished data and numerous d-f stations furnished bearing information over considerable periods. The work was coordinated and cleared through the National Bureau of Standards [NBS] to whom the observations were sent.

The ionosphere reports submitted under these several projects were used by the various branches of the Armed Forces. The establishment of channels for reception of incoming data and techniques for processing it led to the development of a service known as the Interservice Radio Propagation Laboratory [IRPL] devoted to prediction and forecasting of h-f radio propagation conditions on a worldwide basis. The advantages of this work to the communications of the Armed Forces during the war are obvious.

6.1 PROJECT C-13

The several projects in Division 13 dealing with the coordination of ionosphere measurements and d-f errors are C-13, 13.2-88, 13.2-90, 13.2-91, 13.2-92, and 13.2-99.

Section IV of the final report on Project C-13[1] will serve to show future investigators in correlating d-f errors and ionosphere conditions what was attempted and will offer valuable suggestions as to the layout of the job to do this kind of work. Studied in connection with the final report on Project 13.2-92[2] and the bimonthly reports of the IRPL-G series beginning with IRPL-G1, July-August 1944, the early groundwork for the present improved services performed may be ascertained.

Section III of the C-13 report[1] describes the apparatus used. Better and simpler equipment was subsequently developed. Section V indicates the progress of the project with application to radio transmission up to the date of the end of the project. Section II summarizes types of normal and abnormal ionosphere and field-intensity characteristics observed and shows some of these in the form of graphs and of continuous records of relative field intensities over certain propagation paths. The original tabulations and records are on file at NBS and the cooperating laboratories.

6.2 PROJECT 13.2-92

At the termination of Project C-13 a new project, 13.2-92, was instituted. The work accomplished under this project is as follows.

The correlation of d-f errors and causative ionosphere conditions was carried out by five cooperating laboratories located in Washington, D. C., Alaska, California, Puerto Rico, and Massachusetts.

The d-f measurements at all the laboratories were made with the Navy type DAB spaced-loop direction finder. This and the other equipment employed are described in the final report on Project 13.2-92.[2] Measurements were made on a large number of stations distributed in azimuth, distance, and frequency. The results obtained on approximately thirty representative stations dealt with in the report show relationships of bearing errors, field intensities, maximum usable frequencies, and skip distances, geomagnetic disturbances, absorp-

tion, and transmitter antenna directivity. Mention is made of effects of sporadic-E, scattering, and ionosphere disturbances.

The results demonstrated that deviations, often in excess of 50°, occur in transmissions received at the NBS d-f site from stations located in England and Germany. The influence of auroral absorption zone on bearing accuracy over these paths is analyzed and indicates that the steep gradient of absorption between paths passing near and through the zone reasonably accounts for the effects on low operating frequencies. On high operating frequencies, the dropping of the calculated maximum usable frequency for the path below the value of the operating frequency seemed to account fairly well for the large deviations.

Correlation was found between bearing errors and field intensities, the large errors occurring when field intensities are relatively weak. Considerable evidence that large errors might be predicted at times when the maximum usable frequencies fell below the operating frequency was also obtained. Simultaneous occurrences of large errors and severe geomagnetic disturbances were observed on the Berlin-Sterling (D.C.) path and on the Daventry-Sterling path. Only slight evidence of geomagnetic disturbances on bearing accuracy for paths other than those passing near or through the auroral zone was discovered.

The program was considered sufficiently well under way at the end of the project to enable its being taken over by IRPL. Thus the sponsorship of the project by NDRC ended June 30, 1944.

6.3 PROJECT 13.2-88

The final report[3] on this contract[a] with Stanford University deals briefly with choice of site and construction, goes into detail on the calibration and adjustment of the DAB direction finder and mentions preliminary conclusions deduced from the results of data observations.

Calibration was accomplished by means of a target transmitter consisting of a small crystal-controlled oscillator in a metal case with a 4-ft vertical antenna. Measurements were made at 30° intervals and at 300, 400, 500, and 600 ft at fixed frequencies ranging between 2.00 and 17.32 mc. Errors in bands I and II were bad. It was found possible to minimize these errors by redistributing the loop inductances. The error in every case was taken as the difference between the true bearing and the mean of the direct and reciprocal bearings measured by the DAB.

Beginning March 1, 1944, after a preliminary period of training, regular observations were begun on stations in areas suggested by NBS. These stations were in Alaska, Russia, Mexico, Hawaii, Japan, China, and Australia. Data were recorded on weekly summary forms and copies sent to NBS.

For the most part, large deviations were observed to occur during periods when the maximum usable frequency for the path was below the operating frequency of the transmission being observed. However, exceptions to this were noted, especially over multi-hop paths in the Pacific area.

The correlation of bearing deviations with field intensity was good, in that nearly all large deviations were accompanied by correspondingly low field intensities, although the converse was not always true.

6.4 PROJECT 13.2-90

The primary object of this project[b] was to set up a Navy DAB unit at a site appropriate for proximal d-f observations, as a means of studying ionospheric and radio transmission factors of importance in the deviation of long distance d-f bearings. Actual operation of the equipment was carried out under contract between the University of Puerto Rico and IRPL. The final report[4] gives the methods of calibration employed, the means by which the lower-frequency bands were made to have smaller positive errors than they originally had, namely, by readjusting the loop inductances as was done under Project 13.2-88.

[a] Contract No. OEMsr-1122, Stanford University.

[b] Contract No. OEMsr-1101, University of Puerto Rico.

6.5 PROJECT 13.2-91

Work on this project[c] was carried out in Alaska, where the Department of Terrestrial Magnetism, Carnegie Institution of Washington, set up a Navy DAB-3 direction finder near the University of Alaska. Direction finder measurements were made on a 24-hour basis[1] in April 1944 and continued until the project was taken over in July by IRPL. Among the accomplishments were the plotting of mean hourly deviations from true bearings of the stations observed, and production of scatter diagrams of (1) mean diurnal bearing deviations versus mean diurnal geomagnetic K-figure, and (2) mean diurnal bearings of observed stations for a mean diurnal geomagnetic K-figure of 30 (considered normal) on polar coordinates, showing the direction of their deviation from the true bearing.

Primary conclusions from these analyses, which were continued after the project came under IRPL, was that bearing deviations seemed roughly to go in the direction of a north-south line with increasing geomagnetic K-figure. In general it seemed that preliminary results strongly confirmed the desirability of evaluating directional bearings in the light of radio wave propagation characteristics, but did not show much promise of isolating systematic trends to permit application of predetermined correction factors for general use.

[c] OEMsr-1151, Carnegie Institution of Washington

6.6 PROJECT 13.2-99

As in the other projects of this series, a DAB direction finder was set up, this one near Cambridge, Massachusetts, by Harvard University, and bearings were taken on the required stations, the data[8] being submitted to the Bureau of Standards. In addition plans were prepared for conducting sweep-frequency ionospheric observations by automatic equipment. This involved the construction of the necessary equipment. This project[d] was taken over by IRPL at the termination of the NDRC contract.

[d] Contract No. OEMsr-1252, Harvard University.

Chapter 7

MISCELLANEOUS DIRECTION-FINDER RESEARCH

Several projects under the supervision of Division 13 carried out certain technical work on direction finding which is not conveniently or logically placed in one of the other chapters of this volume. This work is summarized here, to complete the record.

7.1 TESTS ON DIRECTION-FINDER SYSTEMS—PROJECT 13-110

In July 1945 Section 13.1 of Division 13 set forth the general thesis that standards for d-f systems should be worked out so that the adoption of suitable standardized procedure (of testing d-f systems) will result in simplification of correlation of data when two or more direction finders are to be compared. At the time, the Services were employing a wide variety of direction finders, differing in respect to their collector systems, bearing indicators, and methods of resolving bearing information. The frequency band coverage included all communication frequencies from very low through ultra-high frequencies.

A conference of representatives of the various Service laboratories with members of Division 13, NDRC, set up proposed test procedures. Central Communications Research, Cruft Laboratory, Harvard University, was assigned the task[a] of investigating the practicability of the proposed procedures by applying them to existing d-f systems and, at the close of the war with Japan, four Army units, SCR-502, SCR-503, SCR-551 and a developmental model of CRD-2, and a Navy DAB installation had been set up by the laboratory staff.[1] Some measurements had been made and arrangements were completed for continuing the work actively under a Navy contract.

7.2 SURVEY OF AIRBORNE DIRECTION FINDERS—PROJECT 13-109

A survey of existing airborne d-f systems for high frequency, very high frequency, and ultra-high frequency was conducted under Project 13-109 and the revised final report[2] gives the result of this investigation together with recommendations for future work. The report gives frequency range, present status, type of indication, type of antenna, weight wind drag, power consumption, and special features for the following direction finders: homing equipment; RC-138-T1, AN/ARA-8, AN/APD-1, M-3100, and C-1900; manually rotatable AN/APA-24; automatic direction finders DBH, DBA, CXGJ-2, CXGJ-5, CXHT, CXHM, CXGG-2, CXGL-2, M-2300, M-3000, and M-4500. These instruments cover the frequency range from 0.25 to 5,000 mc.

Recommendations for Future Work

At the time the report was written no existing equipments covered the lower portion of the radio spectrum. Two pending developments included frequencies just above 2 mc, the DBA and DBH, but these were designed primarily for shipboard installation, and it was only the expressed need for airborne automatic direction finders for high frequencies which prompted the Naval Research Laboratory to suggest that these equipments might be satisfactory for aircraft installation.

Previous experiments in airborne h-f direction finding, as well as knowledge of the reradiation characteristics of aircraft discouraged the installation of h-f direction finders with 360° bearing indication as airborne equipment. The likelihood that the DBA or DBH installed on aircraft will give good accuracy was not promising.

Homing-type direction finders operated satisfactorily at any frequency on an aircraft, yet this survey reveals no homing direction finder existing or under development for frequencies below 18 mc. This was probably not due to electrical difficulties in the design of such equipment, but to the fact that the Service found major difficulties in using homing direction finders for obtaining bearings on

[a] Contract No. OEMsr-1441.

distant signals. To take a bearing with a homing direction finder, the plane's heading had to be varied, and at the time the bearing was taken, the plane's heading and position had to be recorded. This process was considered unduly confusing for the navigator. When the homing direction finder was employed to home to within visual sight of a transmitter, this difficulty did not arise, and it was for such purposes that the homing direction finders in this list were probably intended.

Even in the lower very high frequencies, it was expected that the aircraft structure would cause serious errors in any direction finder which provided indication for 360°. It was for this reason that the accuracy of the CXGJ-2, CXGJ-5, and CXHM was doubtful in the lower portion of their frequency range.

From 100 mc upward, gaps in coverage by automatic direction finders appeared to be due only to lack of sufficient Service interest in the past. Upon completion of developments under way, the only frequencies between 100 and 5,000 mc not covered by airborne automatic direction finders would be from 160 to 225 mc, and the manually rotatable C-2100 covered this. There was a need for an airborne unit to cover 100 to 156 mc which does not have the large wind drag of the CXGG-2.

Above 5,000 mc there appeared to be no development problems peculiar to airborne equipment, and any system which would operate on land or on shipboard could be adapted for airborne installation.

It appeared that new approaches were worth consideration for developing automatic direction finders in the frequency band from 2 to 30 mc, where there was need for equipment which could take bearings on communications transmitters without the difficulties inherent in homing direction finding.

From 30 to 100 mc, there was a similar though somewhat less pressing demand for automatic direction finders. Provided the CXGJ-2 proved satisfactory, research was indicated only in the h-f band.

PART II

APPARATUS DESIGN

Chapter 8

U-H-F RADIO-SONDE DIRECTION FINDER

Development of a simple direction finder for observing the flight of meteorological balloons.[1] Using an Adcock antenna and a single-dipole antenna system with a corner-type reflector mounted on a tripod, an accuracy of ¼° in determination of azimuth and from 0° to a few degrees in elevation was attained when measuring the direction of balloon transmitters operating on 183 mc. Gold plating the reflector wires improved the shielding of the reflector materially.

8.1 OBJECT

To meet the need for a simple and dependable method for observing the flight of meteorological balloons under any and all weather conditions, a simple, easily portable radio d-f equipment was developed.[a] The instrument measures both the azimuthal and the vertical bearings of a small radio transmitter sent aloft on balloons, thus avoiding the problems incidental to the maintenance and synchronization of two ground stations which would be necessary if only the azimuthal bearings were observed.

8.2 APPARATUS

The transmitters employed as a source of signals for the experiments are the type used in radio-sondes. They transmit a vertically polarized wave signal at 183 mc.

The radio direction finder is comprised of an Adcock antenna for measuring azimuthal angles and a single dipole antenna for measuring vertical angles shielded with a corner-type reflector to make the dipole free from the effects of reflected waves from the ground. The reflector system is in fact a secondary radiating system which, when placed in an electromagnetic field, produces a secondary radiation field such that, when properly oriented and placed with respect to the dipole, it neutralizes the effect of the original field at the dipole.

A simple sketch of the instrument is shown in Figure 1. Referring to the figure, the elements marked 1 and 1′, which are self supporting rods of duralumin or other suitable material are each $\lambda/4$ long. Rods 1 and 1′ are coaxially supported, with their adjacent ends spaced approximately 1 cm apart, by insulating supports 2, which are in turn supported by the tubular spacer 3 so as to maintain the rods 1 and 1′ in a plane normal to axis X-X' with the pairs of rods parallel and spaced $\lambda/2$. Rods 1 and 1′ have their inner ends connected together respectively by the line 4. This assembly is a directional antenna of the Adcock type, and is used to determine the azimuth of the incoming wave in a manner to be described later.

Rods 5, similar to rods 1 and 1′, are similarly coaxially supported by insulator 6. Rods 5, constituting a dipole antenna, feed the line 7. The dipole 5-5 together with the shield 8 constitute the antenna assembly for the determination of the vertical angle of incidence of the incoming wave.

The shield 8 is of the corner-reflector type which shields the dipole from reflected waves from ground without impairing the receptive and directive characteristics of the dipole in its reception of the direct waves from the transmitter.

The reflector wires 9, approximately 0.6λ in length are supported so as to be mutually parallel, and at the same time parallel with the dipole 5-5 in two planes, whose intersection is the line D-D'. The included angle between the planes ABC is 60°. The dipole 5-5 lies in a plane which bisects angle ABC at a focal distance p of slightly greater than $\lambda/2$ from the line D-D' and parallel thereto. The reflector wires should not be more than $\lambda/60$ apart, and preferably closer.

[a] Project C-33, Contract No. OEMsr-217, California Institute of Technology. It is understood that a radio-sonde d-f system developed independently by the Air Forces made it unnecessary for the Signal Corps to do any more work on the instrument developed under this project.

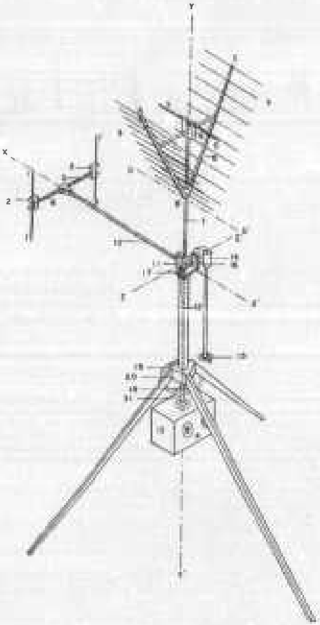


FIGURE 1. Sketch of d-f instrument.

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The two antennas are connected through their feed lines 7 and 10 to a d-p, d-t switch, of suitable design for the frequencies employed, permitting connection at will of either antenna to the line 12 which feeds the receiver 13. Line 10 connects to the electrical midpoint, which is also the geometrical midpoint if carefully constructed, of line 4. Lines 4, 7, 10, and 12 are made with two parallel No. 18 copper wires separated by victron spacers.

The two antenna systems are so mounted as to be rigidly held in fixed positions relative to each other and constitute the directional antenna assembly. The axes *X-X'*, *Y-Y'*, and *Z-Z'* intersect at angles of 90° each with the other. The dipole 5-5 is parallel to axis *X-X'*. Lines 7, 10, and 12 have a length of λ.

The directional antenna assembly may be rotated about axes *Y-Y'* and *Z-Z'*. One means of turning and controlling the rotation of the assembly about axis *Z-Z'* is illustrated where the worm reduction gear 14 is turned by means of the hand wheel 15. Rotation about axis *Y-Y'* may be produced manually or by some mechanical device. The angular positions due to rotation about axes *Y-Y'* and *Z-Z'* may be indicated and measured by any suitable system. The simple device shown in the drawing consists of a graduated quadrant 16 and fixed pointer 17. The fixed graduated circle 18 and its associated pointer, 19, indicate and measure angular rotation about the *Z-Z'* and *Y-Y'* axes respectively.

The complete unit is shown mounted upon a tripod, 20, so that the receiver, 13, is one wavelength above ground. For best results the axis *Z-Z'* should preferably be more than two wavelengths above ground.

The superheterodyne receiver 13 has an output meter to indicate the signal intensity and a pair of earphones for audible indication. It must be well shielded to eliminate stray pickup. The receiver is supported by a tubular support 21, so that it rotates with the antenna assembly as an integral unit with it, and definite advantages result since it makes better shielding possible and there is no possibility of the characteristics of the transmission lines being altered even though the assembly is continuously rotated in the same direction. In this way the relative position of the operator with respect to the two antenna systems will remain unchanged during operation thus eliminating any error in the measurements caused by the changing position of the operator with respect to the antenna system.

RESULTS

Various types of reflector systems were tested as shields of a dipole antenna from ground-reflected waves when used to measure

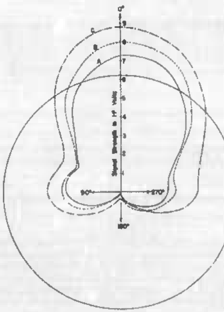


FIGURE 2. Azimuthal response of $\lambda/2$ antenna with 60° corner reflector spaced $\lambda/60$.

the elevation angles of incoming electromagnetic waves. A single rod reflector, reflector and director combination, cylindrically parabolic sheet or wire reflectors, cylindrical sheet or wire reflectors, and corner reflectors were examined.

The corner reflector in conjunction with a simple $\lambda/2$ dipole was found to be most satisfactory for the purpose. Figure 2 is a polar diagram of the strength of the received signal,

in terms of the i-f voltage on the plate of the last i-f stage of the receiver, for various horizontal angular settings of the shield. The circle

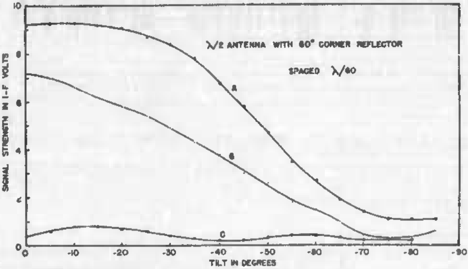


FIGURE 3. Response characteristics with antenna tilted toward transmitter.

represents the uniform signal received without any shield. Maximum shielding is shown where the open side of the shield is 180° from the transmitter. In this setup the λ/2 antenna and the reflector remained vertical, and the reflector was rotated about the antenna. The curves show the results for different focal distances p at a spacing of λ/60 between reflector wires.

It was found that a focal distance of 86 cm gave the best results with a good ratio of shielding to gain as the reflector was swung through 180°. Representative curves A, B, and C show the response for p's of 85, 86, and 100 cm respectively.

Figure 3 shows the results of tilting the λ/2 antenna with and without the shield, toward a stationary transmitter located on Mt. Wilson (about 7 miles away) about a horizontal axis. Curve B is the response of the antenna alone without any shield. Curves A and C are with the shield in place facing toward the transmitter, and opposite to it, respectively. These positions correspond to the respective angles of 0° and 180° in the polar diagram, Figure 2. Curve C shows that the shielding is quite effective and fairly uniform.

In Figure 4 are shown respons curves indi-

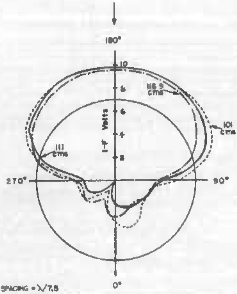


FIGURE 4. Azimuthal response of dipole with 60° corner reflector made of wires of different lengths.

cating the effect of the length of the wire elements of the reflector at λ/7.5 spacing on its shielding properties. Because of the congestion of the curves near the zero-angle region, only representative curves are drawn in the figure. It is seen that best shielding occurs at the length of 0.65λ. This value was used as the optimum length of the wire elements in the later experiments on the wire spacings of the reflector.

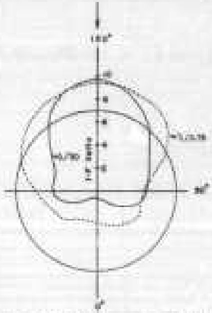


FIGURE 5. Relation between spacing between wires in corner reflector and azimuthal response.

cock antenna and other metal supports of the instrument.

Experiments made with an incoming radio wave emitted from a transmitter at Mt. Wilson at a vertical angle of 7¾° showed that without the reflector the deviation from the true direction is over 22°, while with a reflector of λ/30 spacing the deviation reduces to about 1° (see Figure 7). From Figure 7 it is seen that for a λ/60 spacing the null point is much sharper. It is to be remembered that at this

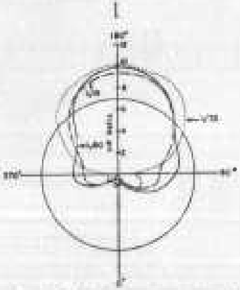


FIGURE 6. Relation between wire spacing and azimuthal response for spacing values not shown on Figure 5.

Figures 5 and 3 show the effect of wire spacing on the shielding properties of the reflector. It is seen that the closer the spacing of the wire elements, the better is the shielding, although it is not too critical when the spacing is smaller than λ/7.5. Nonetheless, λ/60 spacing seems to be the best in the group.

These tests were all made close to ground. It was later found that the results thus obtained do not quite hold when the corner reflector is mounted on the instrument at a height of 3λ above ground, and in the vicinity of the Ad-

grazing angle of 7¾°, the intensity of the reflected wave from the ground is extremely strong. This suggested the necessity of further decreasing the spacing and the results shown in Figure 8 using λ/120 and λ/240 spacing are quite satisfactory. In both cases the deviation is only ¼° which is of the same order of magnitude as the experimental error of the instrument. It is also seen from Figures 7 and 8 that the small humps which appear in the case of larger spacings are smoothed out in the case of λ/120 and λ/240 spacings.

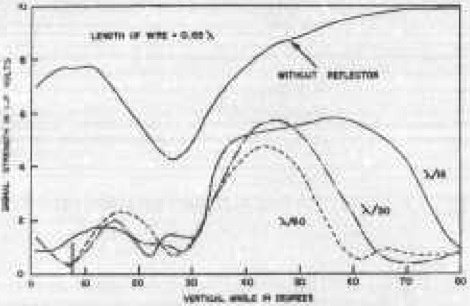


FIGURE 7. Vertical response characteristics of 1/2 dipole with 60° corner reflector with different spacing between reflector wires.

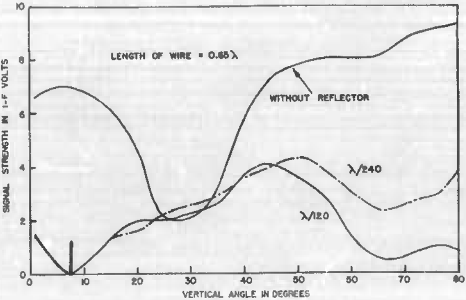


FIGURE 8. Continuation of data shown in Figure 7 with more wires in reflector.

Use of Copper Screening as Shield

For the λ/60 spacing there were about 120 wires which had to be individually fastened into proper place with the right spacing, and for a λ/240 spacing there were 480 wires fastened on to the reflector frame. Some difficulties were experienced in putting on and changing all these wires on a light wooden frame with the spacing so close and yet without the wires touching each other, when they had to be fastened on to the reflector frame which is about 20 ft above ground. The λ/240 spacing is already so close that further decreasing of the spacing is impractical in the present method of mounting. In view of this fact shielding properties of fine copper wire screen were experimented. Figure 9 shows the response curves of the reflector using various numbers of pieces of copper wire screen as the reflecting elements.

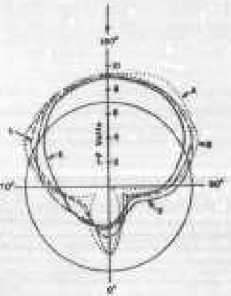


FIGURE 9. Azimuthal response of antenna with brass screen reflector. Curve A, whole sheet of screen; curve B, screen cut crosswise into 4 pieces; curve C, screen in 8 pieces; curve D, screen in 12 pieces; curve E, screen in 16 pieces.

It can be seen from Figure 9 that the more the wire screen is cut, the better is the shielding. This shows that the presence of the horizontal members of the wire screen decreases the effect of shielding.

Different sizes of wires and tubings ranging from No. 32 wire to ¼-inch tubing were tried as the reflecting elements and no appreciable difference in the efficiency of the shielding property of the reflector was observed.

Determination of Azimuthal and Elevation Angles of an Incoming Wave

Directional measurements were made on incoming waves emitted either from a stationary transmitter located on top of Mt. Wilson 7 miles away, or from a transmitter sent aloft on meteorological balloons.

Table 1 shows the results of azimuthal measurements with the Adcock antenna made in an open field with the transmitter supported by a captive balloon. This illustrates the independence of the azimuthal measurements from the vertical incidence of the incoming wave. The last three low-angle measurements were made at different times, at different locations, on the Mt. Wilson transmitter. The accuracy obtained in the azimuthal measurements is within ¼°.

TABLE 1. Azimuthal angles measured by Adcock antenna.

Observations made by Adcock antenna, in degrees	Visual observation, in degrees	Vertical angle (at time azimuthal observation was made), in degrees
88	88	31
87½	88	32½
88	88	29
88	88	20
7	7½	45
9	9½	42
½	½	45
29	29.9	25½
7 (Mt. Wilson)	7.3	8½
½ (Mt. Wilson)	0	7½
0 (Mt. Wilson)	0	7½

Repeated measurements made on the elevation angles of the direction of the incoming wave emitted from a transmitter on top of Mt. Wilson are within ¼° from the true direction

(whose true elevation angle is 7½°). Observations were also made on transmitters sent aloft on captive balloons at various altitudes and elevation angles. The results of these observations made at various times are shown in Table 2.

TABLE 2. Measurements of elevation angles by a dipole antenna with corner reflector.

Determined by dipole antenna, in degrees	Determined visually, in degrees
32	32
28	28
24	21½
58	58
50	50
20½	20
32	32
49	49
58	57½
65½	64½
66½	66
71	69
68	68½
67½	68½
77½	77
51	50½
35½	35
35	34½
33½	33¾
25½	25½
26½	29½
30½	31½
28	29
28	27½
28¾	25½
30½	31½
31	28½
30½	29
29¾	29½
29½	28½
31	34
31	33½
32½	33
33	36½
32½	38½
32½	37¾
40	34½
40¾	36½
41½	39½
41½	39½
41¾	36½
41¾	40¾
41½	39½
40	38½
39½	38
40½	35
43	42½
43¾	43½
43½	44½
43¼	43½
42¼	39¼
43¼	43¼
42¾	40¼

Table 3 shows the measurements made on the elevation angles when the reflector was removed. This demonstrates the great deviation

TABLE 3. Measurements of elevation angles without reflector.

Determined by dipole antenna, in degrees	Determined visually, in degrees
3	8
12	16
7	14
19	14
25	31
28	17
30	29
40	32
60	30
68	34
90	76
98	72

in the readings from the true direction caused by ground-reflected waves.

While some of the results of the elevation angles obtained by using the direction finder are very good and agree within ¼° with the readings obtained visually, there are readings which differ quite appreciably from those measured visually. This is probably due to the fact that the antenna swings badly, changing the plane of polarization of the incoming waves. This in turn affects the magnitude of the emf induced in the receiving antenna and causes the fluctuation in the output of the receiver. When the indicator needle swings badly, it is hard to determine the null point with any accuracy.

EFFECT OF SHIELD OXIDATION

Another factor responsible for the deviations in the readings which sometimes amounted to as much as a few degrees is probably the decrease in the efficiency of the shielding system on account of the formation of a poorly conducting layer on the surface of the copper wire elements. This layer is due to oxidation, caused by the constant exposure of the shielding system to various weather conditions. Since at ultra-high frequencies, practically all the current flowing in the wire is concentrated on the surface of the wire, any contamination of the surface will decrease the efficiency of the wire elements in their secondary radiation. This is

especially important in the present case because the secondary radiation field due to these wires serves to neutralize the effect of the ground-reflected waves at the antenna. Any deterioration in the efficiency of the shielding system would decrease the intensity of the secondary radiation field thus causing the effect of the ground-reflected waves still to be markedly noticeable at the antenna.

neers using fixed and movable target transmitters which were either set up on top of one of the laboratory buildings or carried around by an Army jeep on the field. The results obtained were quite satisfactory. Later a free balloon flight with a buzzer-modulated transmitter was made and the flight lasted about half an hour with a range of approximately 25 miles. The results obtained show that an error in both the

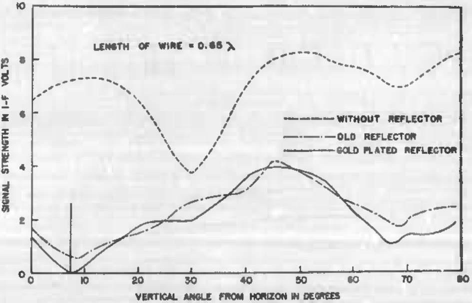


FIGURE 10. Effects of gold plating the reflector.

A new corner-type reflector was made of No. 30 copper wire which was gold plated. Tests made using the new reflector on an incoming wave from Mt. Wilson show a marked improvement over the old reflector whose elements had been badly oxidized. This is shown in Figure 10.

Experiments Made at Fort Monmouth

The direction finder was shipped to Fort Monmouth September 18, 1942, where it was set up on the ground in front of one of the laboratory buildings of the Signal Corps Field Laboratory No. 2 at Eatontown, N. J. Extensive tests were made by the Signal Corps engi-

azimuthal and elevation angles ranged from ½° to 3½°.

During one of the flights, the theodolite observer lost the balloon in a heavy cloud bank, but 25 minutes later the balloon was relocated in the theodolite with the help of the settings obtained by the direction finder.

It is believed that by using proper damping devices to keep the antenna from swinging appreciably during the flight, by properly shielding the transmitter and by further increasing the efficiency of the reflector system such as by decreasing the reflector spacing, etc., the accuracy in the determination of the elevation angles can be greatly increased.

Chapter 9

DEMOUNTABLE SHORT-WAVE DIRECTION FINDER

Development of equipment (SCR-502) giving instantaneous bearings on signals in the region 1.5 to 30 mc, easily transportable in an Army trailer, capable of being erected in a few hours, with wave collectors as free from polarization errors as possible, the cathode-ray oscilloscope patterns which show bearing also giving indication of the quality of the bearings and the condition of operation. As a single-band system (SCR-291) this apparatus was widely used by the Air Transport Service.[1]

9.1 INTRODUCTION

AT THE TIME this project[a] was started, the principal short-wave direction finder in use was the elevated H Adcock system. It was manually rotated by the operator and employed aural null indication. A typical device of this type was the SCR-551-T1. Some preliminary work had been done on a fixed land-station direction finder for these frequencies, designated the DAJ, and made for the Navy. No portable short-wave direction finder was available which would give reasonably accurate bearings under conditions of sky-wave reception, principally because of errors caused by horizontally polarized components of the received signal. In a large number of cases it was impossible to take bearings with existing systems because of the inability of the operator to follow the null mechanically. Tests on an elevated H Adcock showed that its performance was greatly affected by ground conditions and that the order of balance required to secure protection against horizontally polarized waves was beyond all practical limits.

The DAJ equipment showed that it was possible to secure materially improved results by using cathode followers at the antennas and by burying the cables to reduce the effects of current in them to a small degree.

[a] Project C-34, Contract No. OEMsr-263, Federal Telephone and Radio Corporation.

9.2 ACCOMPLISHMENTS

The SCR-502, a two-band system covering 2 to 30 mc and which was produced in quantity as a single-band system for 2 to 10 mc (SCR-291), gave bearing accuracy of 2 per cent on perhaps 75 per cent of the received signals when the apparatus was properly set up, the octantal error corrections made and the operator trained to its use.

Compared to existing systems, the SCR-502 had a great improvement in accuracy, gave the operator the ability to see the nature of the signal and to interpret its probable worth as an indication of bearing, and was reasonably portable since it could be set up in about 2 hours on a suitable level site without the necessity of burying any cables or indulging in airplane calibration procedures. It was comparatively easy of maintenance because of the cathode followers separating the antennas from the connecting cables to the apparatus.

The principal characteristics and advantages of the system are:

1. Bearings can be taken on very short signals. For instance, a good bearing can be taken by an ordinary operator in 2 seconds maximum (including sense finding).

2. The reliability of the bearing is known. The indicator shows whether the signal is mixed with interference or with other signals. It shows if the bearing is steady and reliable, or shifting due to propagation conditions.

3. The signal is audible during the taking of bearings.

4. The accuracy is independent of the frequency within the limits allowed by the quality of the antenna and goniometer designs.

5. Only one receiver is used in a conventional manner. The receiver differs from a standard type in that the input circuit is designed to match the balanced output of the goniometer and the output of the rectified i-f

circuit is connected to the external oscilloscope amplifier.

6. Remote indication is possible. An oscilloscope indicator installed at a distance will indicate the same image as the original. The transmission can be effected on two wire line pairs.

7. Sensitivity of the direction finder for a selectivity of 3 kc and a signal-to-noise ratio of 8 to 1 varies between 6 and 12 μv per meter within the frequency range covered. Under these conditions, a good readable pattern is observed on the indicator so that an inexperienced operator can take a bearing accurate to within 2 per cent and an experienced operator can take a bearing accurate to within 1 per cent. Correction curves of the system installed at a proper site are not larger than ± 4 degrees.

The sensitivity on sense is the same as on direction finding; on downcoming waves with elevation angles up to 30°, the sensitivity is practically as indicated above; half the above figures hold for angles of 45° and one-quarter the values for angles of 60°.

8. Polarization errors as measured by means of a transmitter on a 90-ft tower are no larger than those of the best fixed antennae that could be installed. The monopole antennas are 25 to 50 times less sensitive to horizontally polarized waves than to vertically polarized signals. Standard wave errors are of the order of 1.5 to 3° maximum.

9. Installation of the complete system plus adjustments requires 47 man hours; a trained crew of five men can place the direction finder in operation in two to three hours.

9.3 DESCRIPTION OF THE EQUIPMENT

The direction finder comprises:

1. Two sets of five monopole vertical antennas, only one set to be utilized at a time with the corresponding set of transmission lines. One set covers the range 2 to 10 mc; the high-band collectors covering the range 10 to 30 mc. The two wave collectors differ in the spacing of the receiving elements, and therefore, in the lengths of transmission lines and the dimensions of the ground mats.

Each set of wave collectors comprises four fixed monopole antennas used in conjunction with a crossed-coil goniometer plus a fifth receiving element identical to the four others of the group, installed in the center of this group. This fifth element is used for sense finding. The two wave collectors can be installed in several ways providing a minimum distance of about 60 yards exists between the two.

2. Two remote motor-driven goniometers installed in the field near the antennas and designed to operate in the wave range of each antenna.

3. A receiver covering the entire range in several bands.

4. A CRO indicator.

5. A power-supply system normally designed to be operated from 110 volts alternating current; if this current is not available the equipment can be operated from a storage battery feeding a rotary converter, although the battery drain is high.

6. A remote indicator, all electronic, reproducing at a distance of 7.5 miles the pattern obtained on the local indicator by means of two W110-B type field wire lines.

The receiver, indicator, and power-supply circuits are installed in a trailer. This trailer is arranged to carry all parts of the antennas, reels of cable, ground mats, and electronic equipment.

9.4 PRINCIPLE OF OPERATION

The signals received by a wave collector made up of five receiving elements are converted to balanced outputs by cathode followers, mixed and scanned with a goniometer, the search coil of which is rotated by an electric motor at a constant speed of 30 rps. The signals are then amplified and rectified in the receiver.

The output circuit is arranged so that, in the absence of signal, a continuous current is obtained. The presence of the signal carrier reduces the current, which approaches zero for signals of sufficient amplitude. The output current is applied to a deflection coil system which rotates about the neck of the oscilloscope synchronously with the search coil of the goniometer.

In previous designs the goniometer was installed on the same shaft as the rotating coil, thus avoiding the synchronization problem. In this design, the goniometer is remotely rotated

by a synchronous motor and the synchronism is automatically controlled.

In the absence of signal, the deflected spot traces a circle on the screen. In the presence of signal, the deflection is decreased and the spot tends to come back to the center of the screen. Because of the synchronization with the search coil, a fixed image like a double arrow, which can be sharpened or flattened depending on the amplitude of the carrier current at the detector, is seen on the CRO screen.

9.4.1 Sense Circuit

Sense indication is obtained as follows: the central antenna is connected to the output transformer of the goniometer through a high-frequency line. The resulting cardioid diagram or one intermediate between cardioid and figure eight causes an image as shown in Figure 1 to appear on the screen. To make the sense reading easy, the position of the figure on the screen is rotated 90° by connecting the output of the CRO amplifier to another set of deflecting coils displaced 90° with respect to the normal set. This operation is performed by a relay at the same time as the ... from the central receiving element is applied to obtain a unidirectional diagram.

Figure 1. Sense and azimuth patterns superimposed on CRO indicator. In practice, azimuth pattern is suppressed when taking sense reading.

The CRO indicator used locally (type AS) is highly accurate. The remote indicator (type E) is less accurate but performs satisfactorily the operation desired.

9.4.2 Operation of the Remote Indicator

At the main station a pulse is generated synchronously with the rotation of the type AS indicator. This pulse, sent over one of the telephone lines to the remote indicator, synchronizes an oscillator which is used to produce a circle on the CRO screen. The diameter of the circle is controlled by the coupling tube bias, which is modulated by the output current of the receiver sent on the second telephone line. A pattern similar to that observed on the type AS indicator is produced on the remote indicator.

A voice telephone circuit can be connected to the first line.

The remote indicator is locally supplied with power not required to be exactly of the same frequency as at the main station.

9.4.3 Wave Collectors

Quite a number of antenna systems were studied before the one used in this system was selected. Vertical- and horizontal-spaced loops were compared with the well-known vertical-spaced antenna system and it was found that, for efficiency and sensitivity, the monopole antennas were much better.

However, when using conventional monopole antennas, with or without solid ground mat, and with *direct* shielded crossed connection between each pair of antennas (buried or not), the results obtained with respect to polarization errors were rather poor.

A thorough study of the operation of the monopoles showed that the direct connection between antenna bases through a solid ground mat of small dimensions, or through the shielding of the cross-connecting line was responsible for a very large part of the polarization errors observed.

Therefore, a new system of connection was established, in which the lines approach each

pair of monopoles at an angle of 90° from the position of the cross-connecting lines employed in the old designs.

If, now, the two parallel lines leaving one antenna pair are prolonged to infinity, no parasitic induction will take place at the null point for any polarization of the sky wave.

However, a practical length of these lines before cross connection has been found to be twice the spacing between monopoles. Therefore, the lines and cross connections have the shape of a U lying on the ground.

Study also showed that:

1. Independent ground mats of a radius of about 20 to 30 per cent of the height of the monopole are the most satisfactory. The use of a solid ground mat of small dimensions immediately jeopardizes the quality obtained with the U connections.

2. The connection of the monopoles as explained above, with cables lying on the ground, gives much better results than direct cross connections even though the latter are buried 12 ft in the ground.

The monopoles are coupled to the high-frequency lines through coupling units made of a cathode-follower phase-inverter circuit. In this manner:

1. Only one tube is used to transform the unbalanced input into a balanced output.

2. The circuit efficiently transfers energy from the antenna to the low-impedance line, for any value of the antenna impedance throughout the rather large band required.

3. The same coupling unit is used for frequencies from 1 to 30 mc.

4. There is a loss in the voltage transfer from antenna to line of about 50 per cent, but a gain of power transfer of over 20 db. This is in excess of an ideal transformer.

5. Tube noise is negligible due to the degeneration present. In this respect the circuit works like a transformer, the fluctuation noise still being originated in the first circuit and tube of the receiver.

6. Any length of high-frequency line can be used due to proper impedance matching at the transmitting end.

7. Due to the amount of degeneration and the absence of voltage gain, the transfer is quite stable and the d-f operation can be performed in spite of supply voltage variations as great as 15 per cent.

As a whole, this new means of coupling the antenna to the lines, jointly with the new system of U cross connection between antennas, represents a complete redesigning of the old Adcock antenna.

The use of a remotely rotated goniometer requires the use of only one high-frequency line between the goniometer and the trailer containing the equipment, with consequent reduction in the problem of balancing the electrical lengths of the cables.

9.4.4 Sense Circuit

Determination of sense has always been a delicate feature of the old Adcock systems, especially when the frequency band was large. Amplitude and phase adjustments were needed, and the results were doubtful and required too much time. A new sense circuit has been developed which does not require an extra amplifier, tuned circuit, or phase and amplitude adjustment. In this new circuit the sense monopole antenna is connected to cathode followers each of which is connected through a transmission line to a common dummy goniometer. The secondary of this goniometer is coupled to the secondary of the normal goniometer through a sense relay. The dummy goniometer is electrically equivalent to the rotating goniometer. Its purpose is to introduce within the frequency band a phase shift exactly equal to the phase shift introduced by the normal goniometer. The two transmission lines are of different lengths; the difference being electrically just equal to the spacing between the two monopoles of one directive antenna pair.

The result is that the phase of the sense antenna signal is shifted by exactly 90° through practically all the frequency band that the directive monopoles can cover. Moreover, the amplitude of the signal from the sense antenna is automatically equal to the amplitude of the directive signal for all directions and frequencies without adjustment. This sense circuit avoids any sense-circuit modification of a standard commercial receiver in its use in the d-f system.

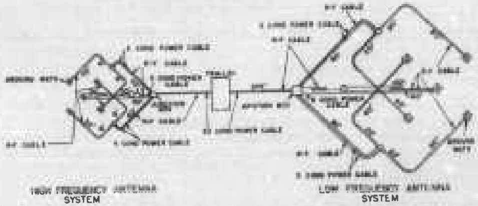


FIGURE 2. Field layout of demountable d-f set SCR-502.

Stand-by reception is provided by cutting off the plate voltage of the coupling units of the four directive monopoles, leaving the sense monopole operating alone.

The cable layout and the area covered by a two-wave d-f system of this type are shown in Figure 2. Other arrangements of the two antenna systems may be employed provided that a minimum distance of 60 yards exists between them. To make the antenna installation easy, a small compass and transit on a tripod are furnished, together with chains of fixed length to determine the spacing between monopoles.

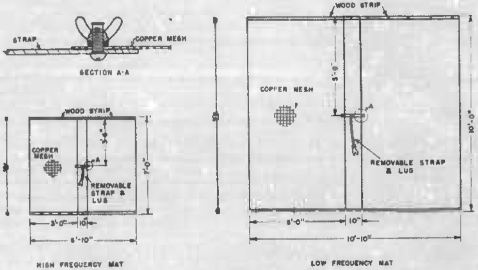


FIGURE 3. Dimensions of ground mats used with two sets of antennas.

The goniometers are of low impedance and without shielding between primary and secondary. They are connected to the receiver without slip rings, through a rotating coupling transformer.

A relay in the coupling unit shorts the antenna to ground in absence of plate current so that any parasitic reception coming from this antenna is avoided, thus permitting a quick check-up of the individual antennas, high-frequency cables, etc.

The ground mats (Figure 3) are made of flexible copper screening material. The dimensions are fairly critical.

The high-frequency lines are flexible solid dielectric in type, balanced, shielded and vinylite covered, made up of two coaxial cables of about 60 ohms impedance. They can be operated in all conditions of humidity or under water.

9.4.5 Goniometer Drive Units

The goniometers (two of which are required for a two-wave collector system) are remote from the antennas and are contained in a goniometer drive unit which also comprises a synchronous motor with synchronization contacts and a junction box for connecting the array to the operating equipment. The relay for connecting the sense antenna line to the primary of the goniometer output transformer is also placed in the goniometer drive unit.

9.4.6 Synchronization System

The synchronous motor rotating the goniometer can take four different positions with respect to the synchronous motor rotating the

FIGURE 4. Functional diagram of synchronizing system to keep indicator and goniometer motors in synchronism.

indicator coils around the CRO tube in the trailer. To avoid a possible ambiguity of 90° an automatic synchronization scheme was developed (see Figure 4). Since the stability of position of the two motors when they are running is better than 1°, the purpose of the synchronizing circuit is only to place the two rotors in the same position without any ambiguity.

To provide for this result, rotating contacts have been placed on the indicator shaft and on the goniometer motor shaft. The contactors are wired in series with one end grounded. The other end of this circuit is connected to the input grid of a two-tube amplifier. This grid is ordinarily biased to cutoff. The contacts are so adjusted that the indicator contactor is closed for about 270° of the rotation and the goniometer contacts close for about 30° of rotation while the indicator contacts are open. In normal synchronized operation the two contactors never close at the same time and thus the input grid of the amplifier is biased to cutoff. Under this condition, grid and cathode of the second of the amplifier tubes are at the same potential and current flows through this tube. Plate current flowing through a resistor produces a voltage drop which is applied to a thyratron as a negative bias preventing this tube from firing.

If, however, the two motors are out of synchronism, there will be a period during the rotation cycle in which the two contactors close together, grounding the grid of the first amplifier permitting it to draw current. Each time the contactors close, a pulse of voltage appears across the plate load resistor and is applied to an RC circuit. After several pulses the potential across the capacitor of the RC circuit reaches the flashing voltage of a neon lamp. Current from the lamp through a resistor puts a negative voltage on the grid of the second amplifier tube, cutting off its plate current and removing the cutoff bias from the thyratron. When the thyratron fires, it opens the power circuit to the goniometer drive motor allowing the motor to slip one pole but stay at synchronous speed. A potentiometer controls the time during which the motor power is interrupted.

These synchronizing cycles will continue until the motors pull into step, usually a matter of from 1 to 3 seconds.

The synchronizing unit also contains a multipole switch which selects the proper circuits for

operation from either the low- or high-frequency array. It switches phase inverter power, goniometer drive, and sense relay circuits, contactor synchronizing pulses and receiver r-f input.

9.4.7 Local Indicator

The local indicator consists of the following components.

1. A synchronous 1,800-rpm motor operating from the 60-cycle 110-volt supply. The motor shaft is provided with the synchronizing contact operation described.
2. A set of magnetic deflection coils mounted in a rotating housing which is also driven by the motor. Provision is made for adjusting the instantaneous angular position of the deflection coils with respect to the motor armature.
3. A 5-in. CRO tube of the electrostatic deflection type which is positioned inside the rotating deflection coils and their housing and whose beam is therefore deflected by the magnetic deflection coils.
4. An optical system which consists of an illuminated scale and a mirror so positioned that the reflection of the scale appears to coincide with the pattern obtained on the cathode-ray oscilloscope.
5. A control box containing circuits and controls for positioning the image on the cathode-ray tube screen and obtaining good focus and correct intensity for easy operation.
6. Housings and brackets for maintaining the mechanical, optical, and electrical parts of the indicator in correct alignment for accurate operation.

9.4.8 Remote Indicator

The remote indicator,[b] an entirely electronically operated unit, displays the same pattern as the local (AS) indicator, and is used in conjunction with the same receiver and goniometer, but does not use any moving parts. It is therefore particularly well adapted for use as a remote indicator.

The circular trace of the CRO spot is obtained by applying to the deflection plates two sinusoidal voltages in phase quadrature generated by a local 30-cycle oscillator the phase of which is synchronized with the goniometer rotation by means of synchronizing pulses. These synchronizing pulses are taken from a rotating contact on the goniometer shaft.

When a signal is being received, the CRO spot is deviated toward the center of the screen by a reduction in amplitude of the 30-cycle voltage. This reduction is accomplished by plate modulation of the tubes which amplify the 30-cycle voltage. The current which causes the plate modulation of the amplifying tubes is obtained from the rectified signal voltage in the receiver.

The phase shift of the indicator pattern for purpose of sense determination is obtained by a 90° phase shift of the synchronizing pulse from the goniometer shaft.

When the remote indicator is used at a distance from the radio receiver, the rectified signal can be transmitted over a standard telephone line into a d-c amplifier of sufficient gain to compensate the attenuation in the line.

In the remote indicator assembly are the following circuits:

1. Synchronizing-time phase shifter. This shifts the phase of the 30-cycle voltage to rotate the position of the arrow on the cathode-ray screen.
2. Oscillator. This is the initial source of the emf for the circular trace. A modified transitron oscillator is employed with output taken from the plate circuit. This form of oscillator is particularly adapted to pulse synchronization over a fairly wide frequency band with the particular virtue that changes in speed of the goniometer will not cause loss of synchronism.
3. Smoothing amplifiers. Since the oscillator wave form is not sinusoidal, a conventional resistance-capacitance low-pass filter in conjunction with an amplifier is necessary to obtain a pure sine wave of sufficient amplitude.
4. Phase splitter. To obtain the circular trace, two 30-cycle voltages in phase quadrature must be available. Accordingly, a phase splitter with semivariable control is incorporated in the circuit design.
5. Dual push-pull modulated amplifiers. To avoid trapezoidal distortion on the cathode-ray

[b] Not supplied with SCR-502.

tube it is necessary to operate with balanced input to the deflection plates. To avoid transients after modulation over a frequency range of 0 to 10 kc, a push-pull d-c amplifier is employed.

6. Modulator. Two tubes in parallel, a low-mu triode and a high-mu pentode, are used to give an arrow pattern which is sharply pointed at the circumference of the circle.

7. Second anode modulator. Because the deflection plates change in potential about 450 volts, and the focus of the cathode-ray tube is dependent upon the difference in the voltage between second anode and deflection plates, the second anode voltage is modulated proportionately with the deflection plate voltage.

8. Audio amplifier. A conventional single-stage amplifier is included so that when the indicator is used as a remote indicator, the operator may hear the radio signal the bearing of which is being indicated.

9. Two conventional power supples are used. Voltage regulation is used in the power supply for the vacuum tube plates but no regulation is employed on the CRO power supply.

Accuracy of the type E indicator is dependent upon purity of wave form in the 30-cycle circuits, freedom from distortion in the modulation and amplifying circuits, and balance in the push-pull circuits. An accuracy of $\pm 3.5°$ was obtained in the laboratory model.

CHOICE OF THE SITE FOR DIRECTION FINDING

Considerable experience was gained during this and other d-f projects on the matter of proper choice of sites for d-f operations. The gist of this experience follows.

The choice of the site is not necessarily the point which would be closest to the transmitter. In fact, long-distance bearings are more accurate than bearings taken at the skip distance. The following simple rules will be found useful in picking a d-f site:

1. Soil.
   a. Must be flat (not more than 1 ft change in elevation in 50 yd).
   b. Must be as homogeneous as possible, capable of supporting grass or plants most of the year, and as wet as possible. Outcroppings of rocks, or large boulders at less than 6 ft from the surface, are undesirable.
   c. A flat prairie, or a cultivated field or pasture is perfectly suitable.
   d. Rocky seashores, rock islands, or rocky hills are unsuitable.
   e. A flat area behind a beach is suitable if the beach ridge meets the following specifications.
2. Obstacles around the direction finder.
   a. From the antenna site the angle between the top of any obstacles and horizon must not exceed 2°. (Obstacles are mountains, hills, trees, houses which mask the view of the horizon.)
   b. An obstacle of 2° cannot be tolerated closer than 200 yd for many antennas.
   c. Long-distance obstacles like mountains can be tolerated if they subtend 5° or less at 5 miles, but it must not be forgotten that they will act as a perfect screen for direct-ray short-wave reception from a transmitter located on the other side of the obstacle.
   d. These obstacles may also affect the intensity and direction of low-angle (10°) sky waves coming from long-distance transmitters.
   e. Such obstacles will not generally affect the reception of sky waves making an angle of 25° or more with respect to ground and coming from medium-distance transmitters.

3. Power and telephone lines. All incoming lines should be laid on the ground and leave the trailer in such a way that they are as remote as possible from the wave collectors. No high-voltage power line can be tolerated at less than half a mile. Such power lines supported by tall steel towers should be at least 1 mile away. The same distance applies to a railroad or a trolley line. On one side of the installation one telephone and/or one low-voltage power line can be tolerated at distances greater than 300 yd from the nearest antenna.

4. Trees. Tall trees (40 to 50 ft) or forest growth must be farther than 300 yd. Small groups of trees not exceeding 20 ft in height can be tolerated if farther than 30 yd from the nearest antenna.

9.6 CALIBRATION

In calibrating the direction finder it is important to operate the target transmitter at a sufficient distance from the collector antenna so the effects of supply and h-f cables are completely avoided. If the calibration is made at too short a distance from the antenna array, errors are observed which are much larger and do not correspond at all to the errors observed with waves coming from distant transmitters.

Figure 5 shows a calibration made on 8 mc, indicating that when the target transmitter is

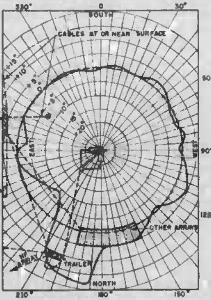


FIGURE 5. Calibration curve of low-frequency array on 8 mc.

near or above the cables, large errors are experienced. In making this calibration the target transmitter was 200 ft from the center of the array at each target position.

To investigate the deviations as a function of the distance of target transmitter, the latter was moved in the direction shown in Figure 6. The diagram on the same drawing indicates for 8 and 3.5 mc how the error varied between 75 ft and 500 ft from the array.

Figure 7 shows a calibration made at 20 mc on the h-f array with the target transmitter on a 75-ft radius. The maximum error observed is 5° in the north direction and is still due to effects of obstacles around the direction finder.

The Great River (Long Island) experimental field where the calibrations were made is far from ideal for such studies with its many antennas and underground cables.

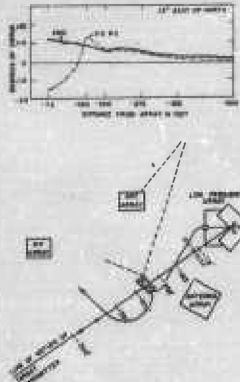


FIGURE 6. Variation of deviation between 75 and 500 ft; 3.5 and 8 mc.

These calibration tests show that, first, calibrations at relatively short distances, where cable effects are magnified by the lack of homogeneity of the field of the target transmitter, have to be completely discarded; secondly, calibrations made at a reasonable distance indicate a vary fair accuracy of the direction finder (small instrumental error) but no con-

stancy in the results because these calibrations are not yet free from effects of obstacles.

Bearings taken on distant stations, 10 miles or farther, show much better accuracy than the best accuracy noted on the calibration curves shown.

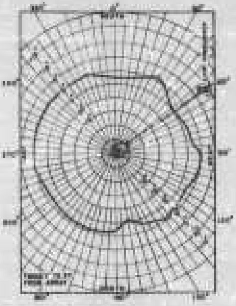


Figure [illegible]. Calibration of high-frequency array [illegible]

About 50 per cent of the bearings taken under any conditions were within 1° and 2° accuracy and only a very small percentage of bearings, less than 2 or 3 per cent, showed errors of more than 10°. These errors were attributed to propagation irregularities or particularly bad interference effects.

[illegible] INTERPRETATION OF PATTERNS

The patterns on the CRO can be read easily by an inexperienced operator. With some training or skill, however, an operator can obtain much information through the interpretation of the patterns which are superimposed upon a fixed scale graduated in degrees from 0 to 360. The equipment is installed and adjusted so that the 0° or 360° point corresponds to a bearing of true north.

In operation, a station is tuned in on the receiver and the gain is adjusted so that a readable pattern is secured. The preferable pattern is that composed of a double arrow (Figure 8A), its points resting on the scale and its central point in the exact center of the scale. The width of the pattern is not important except as this determines the sharpness of the points. Such a signal is usually obtained from a local transmitter having an unkeyed and unmodulated carrier and where reception is well above the noise level of the receiver. The bearing of such a transmitter is easily determined by reading the scale so as to determine the exact point to which the arrow points. Through operation of the sense key, it is possible to determine which of the arrows to read. If the signal is modulated, the point of the arrow is slightly broadened, but the bearing can be read to the same degree of accuracy. The same is not true for a signal which is so weak that it is mixed with receiver noise. The operator is able to determine whether or not the bearing is reliable by observation of the pattern particularly in regard to the following points:

1. Are the arrows sharply pointed? If not, it is possible that the gain of the receiver is not properly adjusted or that the propagation characteristics between the transmitting and receiving aerials are for that moment unfavorable to the determination of the bearing.

2. Are the points of the arrows fixed? If the CRO pattern is shifting, the propagation characteristics are changing with time. This may be due to changing characteristics in the upper atmosphere, or it may be due to reflections from other sources or absorption in the path of the transmission. If the pattern is changing, it will usually be found that not only have the positions of the indicated bearings changed, but also the shape of the arrows is changed. It usually happens that the arrows are broadened and also that the shape of the pattern is distorted from that desired. The oscilloscope gives the actual instantaneous conditions and therefore will indicate the quality of the wave propagation. If the arrow points are sharp and steady, the operator can always be sure that the correct bearing is indicated.

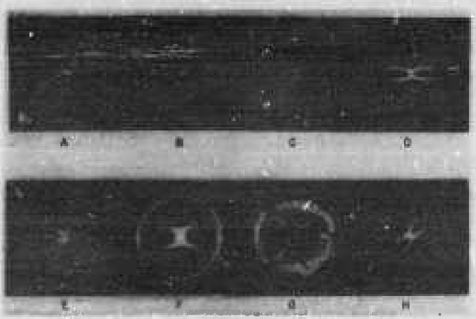


FIGURE 8. Typical patterns illustrating different types of images secured under different conditions of noise and signal.

A. Very strong signal at 90° (or 270°) giving sharp arrow at indications. By manipulation of gain control, arrow may be widened or sharpened but no change of bearing is caused by this operation.

B. Good keyed signal showing circle, smaller than scale circle, when carrier is off air and good indication when carrier is on. Circle can be increased in size by decreasing bias on deflection amplifier.

C. Signal weaker than A and B, but still giving good bearing indication. Arrow points are sharp although maximum signal at center does not cause trace to pull in toward center as in A and B.

D. Much weaker signal in which gain control of receiver has been turned up so high that noise is shown on pattern. Bearing is still readable to within ±3° and possibly higher, since time exposure caused blur of moving arrow points.

E. Modulated signal, strong and without noise, but showing modulation envelope all around outside of pattern. Bearing is not changed by this modulation and is readable to accuracy of ±1°.

F. Time exposure of very weak fading signal. Except for center, entire screen is covered by noise patterns but bearing is still readable to ±5°. Signal is not changing in apparent direction; has little or no polarization error.

G. Time exposure when no signal is being received and with gain control turned to maximum. Although noise appears to have directional characteristics, this would not be visible and is shown here by long time exposure required.

H. Appearance of two keyed signals on same frequency. The one at 90 to 270° is strong and gives good pattern. The one at 42 to 122° is weaker and shows noise envelope as well as slight movement.

Because of the type of indicator employed the operator is able to take bearings on signals of very short duration, or on signals which are rapidly changing at the point of reception. He is provided with a continuous accurate picture of the arrival of the waves and in a short time should be able to take very reliable bearings under conditions which have hitherto made readings impossible.

If a bearing is shifting and also fading, it will usually be found that the nulls are more rounded at one indication than at others. It

will also probably be found that one indication is given with very good nulls. This indication may last only a fraction of a second, but the operator should always remember that the sharpest nulls correspond to the most nearly correct bearing. They always correspond also to the strongest reception during the fading cycle. Round nulls indicate that the bearing is uncertain or in error.

Seldom will no bearing be possible. Bearings will differ only in the degree of accuracy with which they can be read. The most common impossible bearings will be when the noise level is higher than the signal. (Figure 8, F and G.)

A number of patterns are shown in Figure 8 to illustrate the different types of images which may be secured under different conditions of signal and noise.

Chapter 10

DIRECTION FINDING BY IMPROVISED MEANS

A study[a] to determine if effective direction finding could be done by the Armed Forces in war theaters, using only a radio receiver with no special measuring equipment and only such antennas as could easily be improvised in the field. Methods using loops and low horizontal wires were developed. A scheme using low horizontal radial wires each radial having two wires one above the other gave reasonably accurate locations for angles up to about 80° with the horizontal. The text that follows is condensed from the contractor's final report.[1]

10.1 INTRODUCTION

Two general types of antennas were tried, (1) simple loops and (2) various arrangements of low horizontal wires. Table 1 summarizes the results that may be obtained with six schemes briefly described in the table.

The preliminary work indicates that: (1) For strong ground-wave signals, the loop scheme is indicated; (2) for weak ground-wave signals, a rough location may be obtained with a single wire at or near the ground, walked around a central radio receiver. A more accurate location may be obtained using eight radial wires at or near the ground, each wire a wavelength or more in length; (3) for sky waves coming from distances of 150 miles or more, scheme (2) is suitable. For sky waves coming from distances between about 50 and 150 miles, a scheme involving eight radials, each having two wires, one above the other, is indicated; (4) the work on the loop schemes might well be extended to determine methods of locating stations sending sky waves from distances of 150 miles or more. At present, direction but not sense can be determined. Further work on the scheme involving double-wire radials is indicated. A mathematical analysis would be useful to determine a further program of experiments which might lead to refinements and a more accurate determination of the precision of location.

With the loop antenna, the test procedure is to turn the loop for a minimum signal and then reverse the loop by 180° and again find a minimum signal. The best estimate as to station direction is obtained by bisecting the angle between the two minimum-signal positions. A special connection described below permits obtaining the sense on ground-wave transmissions.

All schemes described require an a-m receiver with a beat-frequency oscillator (for producing an audible tone from the carrier) and with the automatic volume control, if any, disabled.

Fairly extensive tests were made of the system listed under Item 6 in Table 1. Part of the tests consisted of locating a mobile station, the direction of which was unknown to the test crew. The distance of the mobile station ranged from 50 to 112 miles. The power into the transmitting antenna was only about 2 watts at 4.8 mc and 6.425 mc. Five tests were made. The average error was about 8° and the maximum error was 22°. The received signal was entirely sky wave.[b] The transmitting antenna consisted of a half-wave horizontal wire about 2 ft above the earth.

10.2 EXPERIMENTAL WORK

Loop Antenna

Two-turn untuned loops, shown in Figure 1, were found to be satisfactory for direction finding on vertically polarized ground waves between 2 and 20 mc. The 2-ft loop is usable in the 2- to 10-mc range and the 1-ft loop in the 5- to 20-mc range. In the 5- to 10-mc range it was found to be more advantageous to use the smaller loop provided the received signal is sufficiently strong.

At first one terminal of the loop was connected to the antenna post of the set and the other terminal was connected to the ground post of the set. This was found unsatisfactory since there was no sharply defined null when the loop was placed broadside to the direction of the transmitting station. (Normally a loop is operated into a balanced circuit.) However, if one terminal of the loop was connected to

[a] Project 13-101, Contract No. OEMsr-1410, Western Electric Co.

[b] The tests were made in the daytime. It is believed that F-layer reflections were involved at both these frequencies. See IRPL-E1, issued September 1944, Figure 15, and TM 11-499, page 45 (both obtainable from Office of the Chief Signal Officer).

the antenna post of the set and the *midpoint* of the loop was connected to the ground post of the set with the other terminal of the loop open, there was a sufficiently defined null when the loop was placed broadside to the direction of the transmitting station. The broadside null for one orientation of the loop was within about 10° of the null for the 180° loop reversal. The average of the nulls gave an indicated direction within a maximum of about 5° of the true direction from 2 to 20 mc when the transmitting station was about one-half mile away over level terrain.

Not only the direction of the station but the sense also may be obtained. The sense is obtained by placing the loop in the plane of maximum signal, a position at right angles to the average of the null, and then touching with the finger or with a pair of pliers held in the hand the free terminal of the loop. With the transmitting station in the direction shown in Figure 1B there will be an increase in the signal. If the transmitting station were in a position 180° from that shown with the loop in the same position, then touching the free end of the loop would result in a reduction of the

TABLE 1. Field of use of improvised direction-finding schemes.

Scheme No.	Kind of antenna (See Figures 1, 2, and 3)	Freq range* in mc	Coverage: Ground wave†	Coverage: Sky wave‡ 70°–80°	Coverage: Sky wave‡ Below 70°	Estimated error§: Ground wave†	Estimated error§: Sky wave 70°–80°	Estimated error§: Sky wave Below 70°
1	Simple 2-turn loop,° midpoint connected to radio-set ground post, one end to antenna post, other end floating	2 to 20	Yes	No	Data incomplete	±10°		
2	Single horizontal wire on ground λ or more¶ in length, walked around central radio receiver	[illegible] to 20	Yes	No	No	±40°		
3	Single horizontal wire λ/4 in length and supported 3 feet high, walked around central radio receiver	2 to 20	Yes	No	No	±30°		
4	Fixed system of four wires on the ground or up to 3 ft in the air, λ or more in length, radially at 90° intervals, with central receiver	2 to 20**	Yes	No	Yes	±23°		±23°
5	Fixed system of eight wires, λ or more in length, radially at 45° intervals, with central receivers	2 to 20**	Yes	No	Yes	±8°		±8°
6	Fixed system of eight double wires, radially at 45° intervals, with central receivers; lower wire of each pair 0.9 to 1.1λ in length‡‡	2 to 8††	Yes	Yes	Yes	±8°	±8° to ±20°	±8°

* [illegible]

† [illegible]ed.

‡ [illegible] to distances of 100 to 150 miles or [illegible]

§ [illegible] schemes there is no "standard-wave error" [illegible]

° [illegible]

¶ [illegible]

** [illegible]

†† [illegible]

‡‡ [illegible] Spread of estimated error between [illegible]

signal. Tests in the 2- to 7-mc range showed that touching the free end of the loop with a wire connected to a vertical antenna produces the same effect as the operator's finger, but the height of the vertical antenna must be adjusted for each frequency.

Three tests (at 3.4925, 4.7975, and 6.425 mc) were made when the transmitting station was within ground-wave range and in directions unknown to the operator. The average bearing error on these three tests was about 10°.

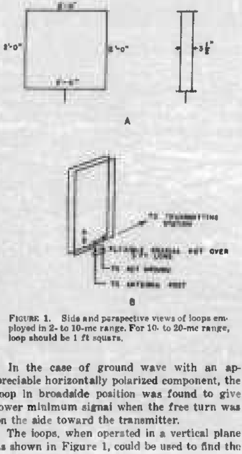


FIGURE 1. Side and perspective views of loops employed in 2- to 10-mc range. For 10- to 20-mc range, loop should be 1 ft square.

In the case of ground wave with an appreciable horizontally polarized component, the loop in broadside position was found to give lower minimum signal when the free turn was on the side toward the transmitter.

The loops, when operated in a vertical plane as shown in Figure 1, could be used to find the direction but not the sense on signals arriving by sky wave from transmitters located at distances exceeding 150 miles. The loop gave a lower minimum signal on sky wave when the free turn was on the side toward the transmitter. However, the results of a few tests made with the loop tilted off vertical and with the free turn on top indicated that a single null could be found which might give both the direction and distance of the transmitter. Time was not available to investigate completely this phase of the problem.

A test was made on WWV at 5 mc where the free end of the larger loop was connected to an eight-wire crowsfoot counterpoise, each toe being about 10 ft long. The radio receiver and associated power supply were not grounded other than through own capacitance to ground, but the connection of the loop to the receiver was through a short length of twisted pair in addition to the coaxial cable. With the loop and WWV in the relative positions shown in Figure 1B a pronounced null was found. When the opposite edge of the loop was pointed at WWV there was a maximum. No broadside null existed; the typical cardioid pattern familiar in the case of a loop combined with a vertical antenna was obtained. There was about 10 db average difference between null and maximum. This scheme, with the same counterpoise, did not work on another station at a different frequency. These tests are mentioned to indicate that the field has not yet been fully explored.

When the loop was connected with its two outer terminals to the antenna post and ground post, respectively, of the receiver, no null of any sort was obtained on WWV or on any other sky-wave signal. This was also true when the loop was connected to a balanced preamplifier, except that in the latter case London and San Francisco stations were found to produce some broadside null.

There is some evidence that the special connection of the loop to the set, described above, tends to reduce its response to the horizontally polarized component of a wave, particularly in the case when the free turn is on the side of the loop toward the oncoming wave.

For sense location the receiver must be placed at ground level, not inside a vehicle, with the loop directly over the set. To facilitate turning the loop any simple mechanical construction may be used. A simple method is to

suspend the loop by a string from a light scaffold directly over the receiver with lead-in consisting of a flexible coaxial cable extending vertically below it to the receiver. The height of the bottom of the loop above the ground should not be over about 3 ft. Army-Navy type No. RG 8/U cable is satisfactory. Twisted pair (for example, W-110B) will not work well for sense indication but may be used for direction indication up to about 7 mc if coaxial cable is unavailable.

Low Horizontal Wires

Low horizontal wires may be used as directional antennas. When the length is $3/4\lambda$ or more, a stronger signal is received when the wave is traveling from the free end of the wire (front) toward the receiver than is received when the wave is traveling in the opposite direction (back) or from the side. The front-to-back ratio will amount to from 5 to 15 db for a wire one wavelength or more in length. For a given length of wire ($>\lambda$) the front-to-back ratio is greater with higher attenuation. That is, there is a greater front-to-back ratio with a wire on the ground than with one supported at some distance above the earth (not over three feet in connection with this work).

The following rule may be used for determining the physical length of *one-wavelength* wire:

$$\text{No. of feet} = \frac{900}{f_{mc}},\ 1\tfrac{1}{2} \text{ to } 3 \text{ feet above ground.}$$

In the clear.

$$\text{No. of feet} = \frac{600}{f_{mc}},\ \text{on short grass.}$$

A $\lambda/4$ horizontal wire is not appreciably directional in itself, but becomes directional when supported in the air and associated with a vertical down-lead. That is, a $\lambda/4$ wire on the ground is not usefully directional as concerns front-to-back ratio, but when raised $1\frac{1}{2}$ to 3 ft in the air, with a vertical down-lead at the receiving end, it becomes directional. In this case, the stronger signal is received when the wave is traveling parallel to the wire from the receiver end of the wire toward the free end. A $\lambda/4$ wire supported up to 3 ft in the air gave poor discrimination in azimuth.

A fixed one-wavelength wire did not give good results on horizontally polarized ground waves.

Calculation and test show that the $3/4\lambda$ wire has a larger front-to-back ratio than the λ wire when the wave approaches the wire from a high vertical angle. However, it has a smaller front-to-back ratio than the λ wire when the wave approaches at a low angle. The simple λ wire gives a good front-to-back ratio on low-angle direction of the wave or on ground wave. This ratio is of the order of 6 db. A wire longer than λ gives a still higher front-to-back ratio. The ratio is affected by attenuation, as noted above, and therefore is affected by ground constants as well as by wire height above ground. The $3/4\lambda$ wire is poor in azimuthal discrimination and therefore is not recommended for use. As discussed later, the $\lambda/4$ wire is useful as a walked wire in ground-wave direction finding but is not successful when used in the fixed-wire radial scheme. The down-lead $1\frac{1}{2}$ ft to 3 ft high also has an effect on the front-to-back ratio of the $3/4\lambda$ wire or λ wire, but the effect may be ignored below about 8 to 10 mc.

Walked Wires

A one-wavelength wire may be used for direction finding on vertically polarized ground waves. The terrain requirements are satisfied by an open field with short grass or weeds of fairly uniform height. The walked wire requires a steady signal, hence it is not useful when fading is present. This limits its use to ground-wave signals.

A full-wavelength wire cut for use on the ground is laid out over grass or weeds and the radio receiver is connected to one end of it. The unknown signal is tuned in, and the wire is walked around to a direction 90° from the first position. At this point the wire is again laid back on the ground and another observation is taken. By progressing 90° at a time, one direction or two directions at about 90° to each other will be found where the signal is fainter than in the other two directions. Further moves, making smaller angular adjustments in the general minimum direction, will disclose a position that gives the faintest signal. In this

position the outer end of the wire is pointing *away* from the "unknown" station.

This scheme is cumbersome at the lower frequencies because of the length of the wire, which becomes difficult to handle and takes a relatively long time to move through an appreciable angle. The scheme works best when a steady carrier is present. With short bursts of carrier, as on average push-to-talk phone operation, the system does not give good results. With c-w or m-c-w telegraph, with steady sending, fair results can be obtained on ground waves.

Quarter-Wavelength Wire. The raised $\lambda/4$ wire with $1\frac{1}{2}$ to 3 ft vertical down-lead is less cumbersome than the full-wave wire. In this case the wire must be held in the air and as nearly parallel to the earth as possible while walking the outer end around. It is preferable to insulate the wire from the walker's hand. When minimum signal is received, then the outer end of the wire is pointing *toward* the unknown station.

This scheme will work with either vertically or horizontally polarized ground waves. Where the wave is horizontally polarized, two positions giving low signals will be found; one with the wire pointing *away* from the station and one with the wire pointing *toward* the station. The one with the wire pointing *toward* the station will be found to be the lowest.

Fixed Multiple Antenna Systems

The use of fixed multiple wires (four or eight), λ or more in length, around a central receiver, in connection with a key which permits rapid switching from one wire to another, may be used for direction finding on ground waves or on sky waves. The wires may be laid on the ground or supported in the air up to a height of 3 ft. Above 10 mc better results are obtained with wires on the ground, unless they are longer than λ. A length of 2λ or more and a height of $1\frac{1}{2}$ ft is satisfactory in the 10- to 20-mc range.

Experiment and calculation showed that where the signal was due to sky waves arriving at the receiving site at an angle greater than about 70° above the horizon, the system was inclined to fail. Fortunately, a failure is revealed in the measurements; hence there is avoided the possibility of taking the "bad" measurements seriously. The failure is revealed by the following symptoms in the test results.

1. The combination of opposite wires giving the greatest front-to-back ratio does not contain the wires which have on them the strongest signals.

2. There may be no consistent front-to-back ratio.

In an example of (1) the station was due north transmitting into a $\lambda/2$ horizontal wire 2 ft high. The greatest front-to-back ratio was SW-NE with SW greater than NE. E and W were greater than SW, and were about equal. As an example of (2), all wires were about equal or they changed back and forth during a fading cycle. At the receiving site near Florham Park, N. J., it was impossible to determine the direction of Floyd Bennett Tower on Long Island at about 7 mc by the above method. The airline distance is about 40 miles. It was possible to determine the direction of WWV in Washington, D. C., at 5 mc. The airline distance is about 180 miles. Successful direction finding was also done by the above method from 6 to about 15 mc on Montreal, Halifax, Toronto, London San Francisco, and on Bound Brook and Wayne, N. J. The last two stations came in on ground wave. Successful direction finding was also done on the project transmitting station when it was about 150 miles distant and transmitting successively at 4.7975 mc and 6.425 mc.

The above method may employ either bare or insulated wires on the ground provided (1) there is no grass or weeds or (2) the grass or weeds are short and uniform where the wires are located. If there are weeds or undergrowth of nonuniform height, the wires should be supported in the air from $1\frac{1}{2}$ to 3 ft high. All the wires in a given layout should be placed at the same height and all should be of the same kind of wire. Test indicates difficulties above 10 to 15 mc if the wires are raised above the earth.

The receiving site should be level. A 5-degree slope, however, will not give trouble. The antennas may be placed in a forest provided raised wires are used ($1\frac{1}{2}$ to 3 ft high) and growth is cut away below the wires to heights

of not over 4 to 6 in. and laterally to a distance of at least 5 ft. Sites near overhead wire lines, wire fences, other antennas or other metallic structures which will distort the field pattern should be avoided. The character of the terrain under all the wires and the growth near them should be approximately the same. For example, it is unwise to put half the wires in the woods and the other half in the open.

When using this method on a fading signal, a phenomenon is present which results in a change of front-to-back ratio during the fading cycle. For example, when listening alternately on wires having about equal signal strength, such as the two wires that are nearly at right angles to the direction of travel of the wave, first one wire and then the other may have a stronger signal. It may be necessary to make the listening comparison for some time to determine which wire gives the stronger signal *most of the time*. The front-to-back ratio on wires most nearly pointing toward and away from the transmitter is consistently in the same direction, but may vary over a range of from 1 to 10 db. This may be quite disconcerting to the operator at first, but it was found that practice in measurements leads to quickness in their interpretation. This practice may be obtained by first testing on known stations.

Double Wires. When the angle of arrival of the sky wave is greater than about 70° above the horizon, (beyond ground-wave range but less than 150 miles) experiment showed that some other scheme than those described above was necessary. It was found that if two wires, one above the other, were put out on each of the eight legs and a slightly different test procedure were used, a great improvement was obtained in short-distance sky-wave direction finding. In this arrangement each leg consisted of a λ wire $1\frac{1}{2}$ ft high and a $\lambda/2$ wire 3 ft high directly above the first wire. In the following discussion the λ wire will be called the low wire and the $\lambda/2$ wire will be called the high wire. The receiver was rapidly connected in succession between two opposite low wires. The associated high wires were connected through and connected to the ground post of the receiver which was also grounded to three ground rods connected in parallel. The other wires of the system remained open and clear.

Now when receiving a high-angle signal which is moving in the direction of the plane containing the wires, the connection of the high wires as noted above *increases* the signal from the low wire pointing toward the station in question and *decreases* the signal from the low wire pointing away from the station. Thus in a case which, without connection of the high wires, would give a front-to-back ratio of unity or less, the connection of the high wires gives a front-to-back ratio greater than unity, i.e., the low wire pointing toward the station (free end toward station) has more signal than the low wire pointing away from the station. To facilitate quick changes, all wires were brought in to a breadboard having Fahnstock terminals and with the switching key screwed to the board. The arrangement is illustrated in Figure 2.

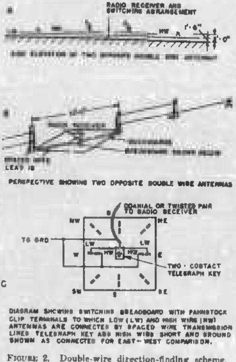


FIGURE 2. Double-wire direction-finding scheme. A. side elevation of two opposite double-wire antennas; B perspective showing double-wire arrangement; C. switching arrangement with key and high wire short connected for east-west comparison.

With this arrangement it was possible to "locate" a mobile station out to 20 miles. There was then a blind ring out to about 60 miles, and from 60 miles on out it was again possible to determine the direction of the mobile station. The extent of the blind ring would vary in different cases depending on frequency, earth constants and on the type of antenna used at the transmitter. The mobile station in question used a low horizontal $\lambda/2$ antenna. Had the mobile station used a whip or other vertical antenna, the ground-wave range undoubtedly would have been extended.

The theory of this system has not yet been completely worked out, but successful use was made of it over a period of about two weeks.

Five tests were made where the mobile station went to locations that were unknown to the measuring crew. The only stipulation was that the distance should be greater than 50 miles. Two frequencies were tested at each location, 4.7975 mc and 6.425 mc. The receiving antennas were arranged to be quickly changed to the right lengths by means of insulators and jumpers. Both frequencies indicated the same direction, excepting for the last test, where it turned out that the mobile transmitter was 50 miles distant. The frequency 6.425 mc gave indeterminate results for this location. Table 2 shows the results. The power into the transmitting antennas was about 2 watts. A Hammarlund radio receiver was used.

TABLE 2. Tests on double-wire system.*

Location	Distance in miles	True bearing in degrees	Measured bearing in degrees
1	63	322	300
2	112	335	330
3	104	15	22.5
4	90	359	355-0
5	50	350	359

* Average error 8°

It is usual to interpolate the bearing between the 45° legs in 15° steps. However, in the case of the reading taken with the transmitter in Location 5, NW was thought to be just slightly stronger than NE and N was stronger than either of these. Hence it was concluded that the station was either due north or very slightly west of north. As can be seen in the table, it was really 1° west of north.

A number of tests were made with the transmitter at 40 miles. This is in the blind ring. In about half the cases tested it was possible to obtain a bearing on the transmitter, most of them at 4.7975 mc. However, in these cases the measured bearing usually came out too large by about one-half the angle between adjacent legs, i.e., 22.5°.

In another set of tests the mobile transmitter was sent, as nearly as possible, due west. On the outward trip tests were made at 20, 40, 80, 120, and 200 miles. On the return trip tests were made at 160, 140, and 80 miles. Errors comparable in magnitude with those shown in Table 2 were obtained for all distances excepting 40 miles. The readings for the latter distance were ambiguous.

On this series of tests a meter was used in the receiver output and it was noted that for the 80-mile transmission, opening the high wire and clearing it from ground reversed the front-to-back ratio from a condition of west wire stronger than east wire by an average of about 4 db, to east wire stronger than west by about 1 db. At 120 miles with the high wire open and clear there was a 0-db (unity) front-to-back ratio, and with the high wire connected there was a front-to-back ratio of about 6 db with west stronger than east. At 200 miles with the high wire open and clear there was a 2- or 3-db front-to-back ratio in the right direction (west greater than east) and with the high wire connected, the front-to-back ratio was about 8 db in the right direction. These comparisons were made at 4.7975 mc. Observations on a broadcast station in Toronto, Canada (about 500 miles), at 6+ mc showed that there was no appreciable difference between front-to-back ratio with the high wire connected or disconnected. The above results suggest that, after further checks of the phenomena, these comparisons might well be used as a criterion of the order of magnitude of the distance of the unknown station.

The high wire and low wire were brought from the terminal stake of each of the antennas to the switching breadboard as spaced transmission lines not over 5 ft long. The wires in

each transmission line were separated by about 3 in. Using twisted pair for lead-in was not tried. See Figure 2B.

The above method permits direction finding on relatively weak signals and on signals which are very close in frequency to other unwanted signals. This is due to the selective action of the ear in being able to identify and concentrate on a tone of a particular pitch in contrast to tones of other pitches or in contrast to noise. A full-wave wire $1\frac{1}{2}$ ft in the air delivers a stronger signal in the 2- to 8-mc range than a 15-ft whip when the wave is moving parallel to the wire from the outer end over average earth.

10.2 GENERAL OPERATING NOTES

In the case of any of the methods described above, a radio receiver outside of a vehicle must be used. The receiver must be placed at ground level for the $\lambda/4$ wire. For sense location, in the case of loop direction finding, the receiver must also be at ground level with the loop directly over the set. For use of any of the full-wave (or more) wire schemes the receiver may be mounted at table top level. The receiver and power supply is not grounded through other than its own capacitance to ground, excepting in the case of the double-wire scheme. In this case, the ground post of the set should be connected to a low-impedance driven ground located near the set. This scheme will work without the driven grounds but it worked better with them.

Power should be supplied by battery and vibropack, both located close to the receiver. A gasoline-driven generator with rubber-covered line on the ground from generator to receiver would probably result in intolerable noise in the antennas. A hand-driven generator probably could be used provided a short length of line between generator and receiver were employed.

10.4 SUPPLEMENTARY TESTS

Tests were made using various other combinations of single- and double-wire antennas, but in the time available no combinations were found that were more satisfactory than the ones described above.

No tests were made above 20 mc. In the range above 30 mc it is likely that the best results would be obtained by use of a reflector-director or other directional antenna. The method would not be applicable to f-m receivers, on account of the limiter action and the lack of a beating oscillator.

10.5 POSSIBLE REFINEMENTS

The fixed-wire methods could be extended in a way that might afford a field of usefulness in other than forward area military direction finding. The refinements would require more apparatus, for example means to connect two antennas through some goniometer coupling device to an oscilloscope for purposes of phase as well as magnitude comparisons.

Modifications can be imagined that would permit rapid direction finding as in many of the existing commercial Adcock systems or spaced-loop systems. Such modifications, applied to the double-wire scheme with eight or more radials, might have particular advantages for high-angle sky-wave direction finding where the Adcock and spaced-loop systems run into difficulties.

The use of the loop with free turns should also be investigated. It is possible that a single loop arranged in this way to suppress the horizontally polarized component might be used in place of two spaced loops.

10.6 THEORETICAL DEVELOPMENT

The proportions of the antennas described in this report were determined by cut and try methods. The general theory of loops and long-wire antennas was used as a qualitative guide. It appeared that this method of attacking the problem would be more efficient than proportioning the antenna structures on the basis of predetermined calculation formulas. The experimental work has given clues to the physical approximations which are justified and which are essential to obtaining reasonably compact calculation formulas.

A calculation formula has been worked out for a simple case. This formula will be discussed together with the physical approximations leading up to it. The method of analysis could be extended to cover the more complicated cases. Figure 3 shows the calculation formula for the case of two collinear horizontal wires at or near the ground and free from ground at both ends. The open-circuit voltages to ground at the inner ends of the two wires were calculated. If the antenna terminal of a grounded radio-receiving set (which in itself does not pick up any voltage from the radio field) is switched from one wire to the other, the relative amplitude of the sounds heard at the output of the receiving set will be proportional to the relative magnitudes of the open-circuit voltages. Therefore, the magnitude of the ratio of these two calculated voltages gives the observed front-to-back ratio. It is assumed, of course, that the radio set does not have automatic volume control.

The radio set does, in itself pick up voltage from the radio field if the down-lead is regarded as a part of the set. Experiments indicated that if the down-lead were not more than about 1/40 of the length of the horizontal wire, the effect was unimportant. The theory could be extended to include the effect of voltages introduced in the down-leads.

The experiments also indicated that it was satisfactory to assume that each horizontal wire was a ground-return transmission line with uniformly distributed constants and 100 per cent reflections at the open ends. Since the wires are near the ground, the transmission line may be regarded as having uniformly distributed resistance. This is caused largely by losses in the ground; the radiation resistance is relatively unimportant. The *magnitude* of the voltage induced per unit length in the horizontal wire may be assumed to be the same at all points of the wire. The coupling between the wires forming different radials may be ignored.

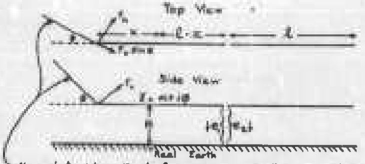
As noted in Figure 4, the front-to-back ratio would be unity if the transmission line had no attenuation. If the attenuation is large, a very simple expression for front-to-back ratio is obtained.

TABLE 3. Front-to-back ratio for pair of collinear wires at or near the ground.

(Propagation constant of wire ground circuit $= \gamma = \alpha + j\beta$, $\beta = 2\pi/\lambda$, $\lambda = {}^{*}k\lambda_a$, λ_a = wavelength for propagation in air.)

Elevation angle in degrees	Azimuth angle ϕ in degrees	Attenuation in nepers	k	λ_a in meters	Front-to-back ratios for wires of length				
					$\lambda/8$	$\lambda/4$	$\lambda/2$	$3/4\lambda$	λ
10	0	0.32	0.92	50	1.00+	1.01	1.16	1.26	1.34
10		0.64	0.92		1.00+	1.02	1.34	1.54	1.78
10		0.67	0.70†		1.00+	1.02	1.26	1.73	1.56
80		0.32	0.92		1.00+	1.00+	1.03	1.38	1.06
80		0.64	0.92		1.005	1.01	1.05	1.85	1.14
80		0.67	0.70†		1.00+	1.00+	1.04	1.67	1.13
80	0	0.64	0.92	50	1.005	1.01	1.05	1.85	1.14
	30				1.00+	1.00+	1.04	1.69	1.13
	45				1.00+	1.00+	1.04	1.60	1.11
	60				1.00+	1.00+	1.03	1.41	1.11
	75				1.00+	1.00+	1.01	1.20	1.11
	90				1.00	1.00	1.00	1.00	1.00
10	0	0.64	0.92	50	1.00+	1.0[illegible]	1.34	1.54	1.78
	30				1.00+	1.02	1.29	1.62	1.63
	43				1.00+	1.01	1.22	1.75	1.49
	80				1.00+	1.01	1.15	1.99	1.33
	75				1.00+	1.005	1.08	2.12	1.17
	90				1.00	1.00	1.00	1.00	1.00

* Per cent of the speed of a wave in space.
† Wire on the ground, wire-ground speed.



Normal to Wave Front of Arriving Elliptically Polarized Plane Wave

Front to back ratio $= \left|\frac{e_1}{e_2}\right| = 1$ if $\alpha = 0$; $\left|\frac{e_1}{e_2}\right| = \frac{\beta+\beta_1}{\beta-\beta_1}$ if $\epsilon^{-\alpha\ell} \ll 1$

$\gamma = \alpha + j\beta$ = propagation constant per meter of wire-ground transmission line

θ = angle of elevation with respect to the horizontal.

ϕ = angle of azimuth with respect to antenna wire.

β_0 = phase change constant for propagation through air $= \frac{2\pi}{\lambda_0}$ radians/meter.

$\beta_1 = \beta_0 \cos\theta \cos\phi$.

r_v and r_h are vector reflection coefficients for vertically or horizontally polarized plane waves.

$$A = F_v \sin\theta \cos\phi\,(1 - r_v \epsilon^{-2j\beta_0 H \sin\theta}) + F_h \sin\phi\,(1 - r_h \epsilon^{-2j\beta_0 H \sin\theta})$$

$$e_1 = A\int_0^\ell \frac{\epsilon^{-j\beta_1 x}\left[\epsilon^{-\gamma(\ell-x)} - \epsilon^{-\gamma(\ell+x)}\right]}{1-\epsilon^{-2\gamma\ell}}\,dx$$

$$e_2 = A\int_0^\ell \frac{\epsilon^{-j\beta_1 x}\left[-\epsilon^{-\gamma x} + \epsilon^{-\gamma(2\ell-x)}\right]}{1-\epsilon^{-2\gamma\ell}}\,dx$$

$$e_1 = -K\left(\epsilon^{+j\beta_1\ell} - \cosh\gamma\ell - j\frac{\beta_1}{\gamma}\sinh\gamma\ell\right) \doteq -K\left(\epsilon^{+j\beta_1\ell} - \cosh\gamma\ell - \frac{\beta_1}{\beta}\sinh\gamma\ell\right)$$

$$e_2 = K\left(\epsilon^{-j\beta_1\ell} - \cosh\gamma\ell + j\frac{\beta_1}{\gamma}\sinh\gamma\ell\right) \doteq K\left(\epsilon^{-j\beta_1\ell} - \cosh\gamma\ell + \frac{\beta_1}{\beta}\sinh\gamma\ell\right)$$

$$K = \frac{2A\gamma\epsilon^{-(\gamma+j\beta_1)\ell}}{(1-\epsilon^{-2\gamma\ell})(\gamma^2+\beta_1^2)}$$

$$\text{for } \beta\ell = 2\pi,\quad \left|\frac{e_1}{e_2}\right| \doteq \left|\frac{\cos\beta_1\ell - \cosh\alpha\ell - \frac{\beta_1\sinh\alpha\ell}{\beta} + j\sin\beta_1\ell}{\cos\beta_1\ell - \cosh\alpha\ell + \frac{\beta_1\sinh\alpha\ell}{\beta} - j\sin\beta_1\ell}\right|$$

FIGURE 3. Calculation of front-to-back ratio; sky-wave reception.

The formulas may be used for calculating ground-wave front-to-back ratios. In this case

$$A = F_v \tan T \cos \phi$$

where T = the tilt angle of the ground wave. Since

$$\theta = 0$$

$$\beta_1 = \beta_0 \cos \phi .$$

Table 3 gives calculations of front-to-back ratios for various assumptions regarding attenuation and length of the wires. The values of attenuation were chosen as the result of previous calculations and measurements on low horizontal wires. The attenuation per wavelength (of the wire ground circuit) is roughly constant (2:1 variation) over a range of about 2 to 10 mc, for heights from about 2 cm to about 1 meter and over a range of ground conductivities from about 0.001 to 0.01 mho per meter. Values between 0.6 and 0.7 neper per wavelength were chosen for calculation. Since these values seem somewhat on the high side, a value of about one-half as much was also used to indicate the effect of reduced attenuation.

The order of magnitude of the calculated ratios is the same as those measured except that the experiments have a reversed front-to-back ratio for $\lambda/4$ wires 3 ft above the ground. The experiments showed these to be due to voltages induced in the down-leads. Computations of front-to-back ratios were first made for zero azimuth angle, i.e., for collinear wires, one of which points at the transmitter. For such wires the angle $\phi = 0$. For lengths of wire which gave a sizable front-to-back ratio, computations were also made for values of ϕ from 0° to 90°. These correspond to collinear radial wires not pointing at the transmitter. It will be seen from Table 3 that the front-to-back ratio does not shrink to unity rapidly as ϕ increases from zero. In one case the maximum front-to-back ratio is obtained on the pair of wires not pointing toward the transmitter. Such false indications were noted during the experiments when arrangements of wires not of the optimum length were used. Since the measurements of only front-to-back ratios do not give a sensitive indication of the direction of the transmitter, it is necessary to compare open-circuit voltages at the inner ends of the wires which are not in line.

Taking as a reference a wire pointing toward the transmitter and for which $\phi = 0$ and considering other wires of angular displacement $\pm\phi$, an inspection of the formulas shows that the magnitude of the open-circuit voltage for any wire is approximately proportional to:

$$\left|\frac{A}{\beta^2 - \beta_1^2}\right| \left|\epsilon^{+j\beta_1 l} - \cosh \gamma l - \frac{\beta_1}{\beta} \sinh \gamma l\right| .$$

In the above expression, α is neglected in the terms having β in the denominator, i.e., β is substituted for $\alpha + j\beta$. Since $\beta_1 = \beta_0 \cos \phi \cos \theta$ it is a function of ϕ. A is also a function of ϕ.

For ground waves A is proportional to $\cos \phi$. For high-angle sky waves (θ = 70° to 90°), the reflection coefficients r_v and r_h are about equal in phase and magnitude. Assuming them alike and assuming F_v and F_h have the same rms values for a time interval of a few seconds but are related at random as to instantaneous magnitude, A is proportional to

$$\sqrt{\sin^2 \theta \cos^2 \phi + \sin^2 \theta} .$$

For low-angle sky waves, the resultant electric force due to arriving and reflected waves is greater for vertical polarization than for horizontal polarization, that is,

$$v = |1 + r_v \epsilon^{-2j\beta_0 H \sin \theta}|$$

is greater than

$$h = |1 - r_h \epsilon^{-2j\beta_0 H \sin \theta}|$$

For 6 mc, ground conductivity of 0.008 mho per meter and ground dielectric constant of 10, the following values of $|v|$ and $|h|$ result if the wires are 3 ft above ground:

θ	r_v	r_h	$\lvert v \rvert$	$\lvert h \rvert$
10°	0.3/100°	0.96/2°	1.10	0.084
50°	0.66/16°	0.79/9.6°	0.50	0.37
80°	0.725/12.5°	0.74/12°	0.46	0.45

For angles where $|h|$ is appreciably less than $|v|$, the coefficient A is proportional to:

$$\sqrt{|v^2| \sin^2 \theta \cos^2 \phi + |h^2| \sin^2 \theta} .$$

Using these assumptions and approximations, the ratio of the voltage on a reference wire pointing toward the transmitter to the voltage on a wire of angular displacement $+\phi$ may be computed. Results of computations for $\frac{3}{4}\lambda$ and λ (wire-ground speed) 3 ft above the earth are given in Figure 4. A frequency of 6 mc and values of elevation angle θ of 10°, 50°, and 80° were used, α was taken as 0.014 neper per meter (0.64 neper per wavelength).

It will be seen that $\frac{3}{4}\lambda$ wires which give relatively large front-to-back ratios for high-angle sky waves give poor azimuthal discrimination. It will also be seen that λ wires have good azimuthal directivity for low- or medium-angle sky waves.

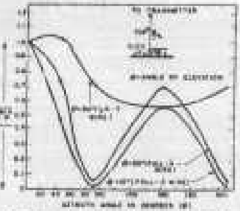


FIGURE 4. Voltage ratios for wires near ground.

Chapter 11

PORTABLE RADIO ASSAULT BEACON

Development of a radio beacon to guide an infantryman to an objective for a distance of 2,500 yd with an accuracy of ±3° using equipment already available. Choice of antenna systems, modulation methods, frequency, solution of key-click troubles are described. The greater part of the contractor's final report[1] on this project is contained in this summary, the chief deletions being descriptions of Army requirements, changes in requirements and in description of methods of aligning antennas in the field.

11.1 INTRODUCTION

At the beginning of this project[a] portable radio assault beacons were in use by the British for the guidance of tanks, the assemblage of paratroopers, etc. The British beacon used two Beverage antennas at right angles. Each antenna was several hundred feet long and was stretched along the ground, usually supported a few inches above the ground.

In this country, some attempt was made to use a crossed-loop antenna system as a beacon for guiding troops to pill boxes, and other targets through fog, smoke, jungle, or at night. The crossed loops, however, gave results which were inferior to the British system. This project studied the performance of the British beacon under various conditions of terrain, weather, antenna length, frequency, obstructions, size of antenna wire, angle between the antennas, height of the wires above earth, inequality of antenna currents, and the polarization error under various conditions.

Existing Army radio transmitters and receivers were incorporated into a beacon similar to that of the British but provided with means for steering the defined course over an arc of 15° or 20° to obviate the necessity of laying out the long antennas very accurately.

The original requirements that the course accuracy be ±½° were manifestly impossible of attainment because average site errors in d-f systems are greater than this figure. The requirement was modified to be ±3°. Original instructions that the equipment was to be the best possible was also changed to a request that every effort be made to utilize equipment already available in the field and to make as few changes in this equipment as possible.

These requirements limited the field of study considerably and finally the development was centered around the use of the SCR-536 receiver and SCR-284 transmitter.

[a] Project 13.1-100; Contract No. OEMsr-1261, Raymond M. Wilmotte.

11.1.1 Selection of Type of Beacon

Six types of beacons were considered.

1. Crossed loops set at 45° to the required direction.
2. Crossed loops set at 0° and 90° to the required direction.
3. Crossed Adcock antennas.
4. Spaced antennas.
5. British type using Beverage antenna.
6. Modified British type using Beverage antenna.

The d-f systems most commonly used are the crossed coil and the Adcock system. They have been used very successfully for aircraft navigation. Because of portability requirements, it is clear that the Adcock type of antenna could only be used at very high frequencies. Because of site errors the use of very high frequencies was discarded. No work, therefore, was carried out with Adcock antennas. Work with crossed loops was successful but the results proved to be less satisfactory than with the British and modified British systems described below.

11.2 LOOP ORIENTATION TO ELIMINATE KEY CLICKS

It was soon found that one of the major difficulties was the elimination of key clicks which were likely to be so strong as to seriously reduce the sensitivity of operation. It was suggested that the crossed loops be located at 0° and 90° with respect to the desired direction

instead of at 45° angles with the direction in the center. The current in the 90° loop would have its current reversed in direction to produce the switching of the antenna pattern. The current in the 0° loop would remain constant so that the signal strength of the signal along the desired direction would not change during the switching period. This method was found successful.

11.3 BRITISH AND MODIFIED BRITISH SYSTEMS

The two systems which seemed to give the most promise were the British system and a modification of it. The British system consists of two Beverage antennas set at 45° to the required direction. The antennas are stretched a few inches above the ground and act in a manner very similar to the ordinary crossed-coil system. Like the crossed-coil system, the British beacon suffers from troubles due to key clicks. That problem was not solved for a considerable time, however. Eventually a relay was developed which reduced the key clicks to such low intensity that the British system was found to be accurate and sensitive.

The modified British system was an attempt to eliminate the key clicks in a manner similar to that in which they are eliminated in the crossed-loop system, i.e., by locating the ground antenna at 90° to the required direction and by use of a vertical antenna. This system was found successful and from an operating point of view was almost identical in sensitivity to the British system but in certain respects of installation was somewhat more complicated.

11.4 SELECTION OF MODULATION

As regards modulation the British reported a hand-operated system in which an operator switches from one antenna to the other and simultaneously speaks into a microphone saying "left," "right," etc., as he switches. The listener then judges the relative intensity of the words "left" and "right" and goes to the left or right accordingly until the intensity of the two words appears approximately equal. Some British reports have indicated remarkable accuracy with this system of modulation. Experiments under this project, however, did not show the degree of accuracy claimed in those reports for normal operating conditions. Other forms of modulation systems such as the dot and dash systems were tried but were not found to be as accurate or as sensitive as the results obtained with the standard A and N system used on aircraft radio ranges.

The modulation can be carried out in one of two ways. When the listener is away from the required direction he must hear a difference in intensity between the signals as the transmitting antenna is switched. This difference in intensity may be obtained either by a change in intensity of the radio frequency or a change in the percentage modulation. In practice it would be preferable to change the intensity of modulation because in that case the automatic volume control [AVC] of the receiver could be used to its full extent without decreasing the sensitivity. When the r-f intensity is changed, however, it is essential that the AVC be eliminated, or that the time of the dots and dashes be short compared with the time constant of the a-v-c circuit, or that the intensity of the signal be sufficiently weak so that the AVC does not operate, or operates only partially.

Since it was eventually decided to try to develop a system using equipment available in the field without making any internal modifications, it was clearly not possible to use the system in which the percentage modulation was changed. The system eventually developed was based on the compromise of using a signal which was sufficiently weak so that the AVC of the receiver is only partially operative, thereby permitting the detection of changes in signal intensity. One method of detecting small changes in modulation was discussed but eventually discarded because it would have required careful operation by the infantryman. This method consisted in using a limiter in the receiver, such as is available in f-m receivers, and adjusting the level of the signal at the limiter so that small changes of signal intensity produced a large change in receiver output.

11.5 SELECTION OF FREQUENCY AND POLARIZATION

The original requirement of an accuracy of $\pm\frac{1}{2}°$ was greater than the accuracy normally

obtained with direction finding. Since it was also clear that any known beacon system would suffer from errors similar to d-f errors, it appeared that even though an accuracy of $\pm \frac{1}{2}°$ might be obtained the absolute direction would probably not be known with an accuracy of better than $\pm 2°$ for most conditions, and under some conditions, a considerably greater error might be obtained. A few tests indicated that it was probable that a system could be developed which would give a high degree of sensitivity. Therefore, it became apparent that, for practical purposes, it was essential to keep site errors to a minimum. Originally it had been suggested that frequencies between 40 and 48 mc be used. This suggestion was based largely on experience with radio ranges at airports and because at those frequencies it might be possible to use spaced antennas thereby providing greater sensitivity than would be possible with the comparatively blunt type of directional pattern that a loop or Adcock antenna provides. Conditions in the field, however, were found to be substantially different from the conditions found at airports. It was also decided that the vast amount of information available on d-f errors should be used in deciding which group of frequencies would reduce the site errors to a minimum or at least to practical values. Dr. Smith-Rose indicated from his experience, which also checked with the experience of the contractor, that the very high frequencies would produce greater errors than lower frequencies. He suggested using frequencies around 300 kc and lower, and that if such frequencies could be used it was probable that site errors as low as 1° might be obtained. However, no equipment was available in the field for these low frequencies and the size of equipment for these frequencies was likely to be excessive if reasonable efficiency was to be obtained. Moreover, the Army indicated that these low frequencies might not be available for this use in the field. Smith-Rose pointed out that the errors increased with increase in frequency and reached a minor maximum between 8 and 10 mc for the reason that at those frequencies an average tree is approximately $\lambda/4$ long and by its resonance causes comparatively large errors. He indicated that errors of the order of 2° could be expected in this range and suggested the use of frequencies around 15 mc with an expectation of reducing the average site error by about $\frac{1}{2}°$. The frequency eventually selected was 5 mc because equipment was available in the field at those frequencies.

Originally NDRC indicated an interest in studying the difference in operation between horizontally and vertically polarized waves. The selection of a frequency of 5 mc eliminated the use of horizontally polarized waves, for within a few hundred feet of a horizontally polarized antenna most of the horizontally polarized waves seem to be eliminated and only the remaining vertically polarized portion is received. The suggestion had been made because it was believed that the horizontally polarized waves would be able to travel farther through wooded territory and produce less error than the vertically polarized waves, since trees are mainly vertical. No work was carried out on this angle of the project because of the decision to use a frequency in the h-f band instead of the v-h-f band, and because it was believed impractical for an infantryman to carry a nondirectional horizontally polarized antenna.

11. EXPERIMENTAL RESULTS

Crossed-Loop Beacon

The crossed-loop system is shown in Figure 1. The loop at right angles to the course is connected to the transmitter through a keyer which reverses the polarity of the currents in this loop in accordance with an A and N. It

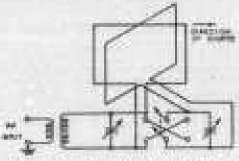


FIGURE 1. [illegible]

was hoped that by performing the switching in that loop which has a null in the direction of the course, the occurrence of key clicks on course would be prevented. To avoid detuning the other loop whenever the keyed loop was disconnected in the process of switching, it was necessary to resonate separately each loop. If the two loops were tuned by a single capacitor, the unkeyed loop was detuned to such an extent when the keyed loop was disconnected during the keying process that the keying was fully reproduced in this loop. The final circuit shown in Figure 1 was more complicated than the British system and the tuning procedure required considerable care.

The field strength obtained with a loop 2 ft square was about ¼ that of the British beacon and averaged about 12 μv at 1½ miles. The width of the course was $\pm 2°$ for a 1-db difference between the A and N signals. Experience with loop antennas amply indicated that no appreciable improvement could be obtained by increasing the size of the loops and it was also felt that larger loops would be undesirable from the standpoint of portability. Because of the low field strength, the greater course width and the greater difficulty of tuning this system, the crossed loops were abandoned. It is possible that the crossed-loop beacon would give satisfactory results at frequencies of the order of 40 to 48 mc and also near 20 mc.

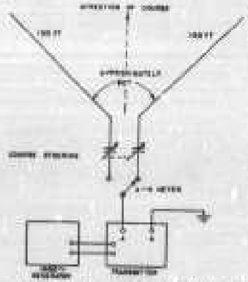


Figure 2. Antenna system used in British aircraft beacon.

The British Type of Beacon

The transmitter used was the SCR-284 which has a maximum output power of 5 watts. The antenna was connected as shown in Figure 2 with a keyer switching from one antenna to the other. The first tests were carried out using the words "left" or "right" spoken in the microphone in accordance with the British method. In the first tests each antenna was about 220 feet long following the British recommendations. It was found, however, that shorter antennas could be used just as effectively and eventually antennas as short as 100 feet were used.

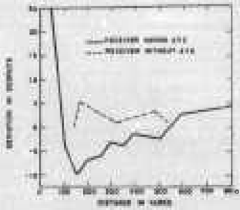


Figure 3. Course of British beacon with voice modulation. Solid line for receiver having AVC; dotted line for receiver without AVC.

Results with "Left-Right" Modulation

Using antennas 220 ft long, tests were made on the width of the course as detected by a nontechnical person who had been given some training in listening to the signals. A Navy RBZ receiver was used. The frequency used was 5.8 mc.

Tests were made with AVC both on and off. The signal was sufficiently weak in most cases that the AVC did not have any large effect.

The results, of which those shown in Figures 3 and 4 are typical, indicate that the course defined in this manner of operation is very wide, ranging up to 8° for medium and long distances and has very poor definition within the first 100 or 150 yd. Although these tests are highly subjective they do give an indication of the bluntness and inadequacies of this type of beacon compared with the requirements of the Army for this equipment. While the course could be followed more accurately by well-trained personnel carefully controlling the input and output levels of the receiver, it was apparent that the required accuracy could not be attained under normal operating conditions.

FIGURE 4. Width of course of British beacon. [illegible]

One reason for the bluntness of the course according to this system was the difficulty of distinguishing differences in loudness between two dissimilar sounds occurring at different times and probably spoken with different degrees of loudness. It was expected, therefore, that considerably increased sensitivity would be obtained by using tone modulation.

RESULTS WITH A-N MODULATION

The use of tone modulation (1,000 cycles) produced a great improvement in the sharpness of the course. An important limiting factor appeared to be the key clicks. It was found difficult to compare accurately the loudness of the A and N signals in the presence of strong key clicks. These key clicks were frequently so much stronger than the tone signals that the observer had difficulty in eliminating from his mind the clicks and concentrating on the tone. The result was often very confusing except to the highly trained personnel. Much work was carried out on the elimination of key clicks. It was found that they could be eliminated or reduced to negligible amounts either by using the modified British system or by a relay of special design.

After the key clicks had been substantially removed it was found that the speed of keying could be increased to 64 characters a minute and still be comfortably read by an untrained observer. Under those conditions the width of the course obtained was about ± 1° for a 1-db difference in signal level but aurally the course width was ±½° to trained observers. The site errors were considerably more than this, being of the order of 2°. The course could be followed with a receiver having AVC such as the SCR-536 and the Navy type RBZ. However, to attain good sharpness of the course with such receivers it was necessary to retract the antenna, slightly detune the receiver, or otherwise maintain the receiver input sufficiently low to minimize the a-v-c action. This was particularly important at the closer distances, say the first 400 yd. It was also found important not to overload the receiver by permitting excessive input voltages, since this could cause an apparent reversal of the A and N quadrants when considerably off course.

FACTORS AFFECTING OPERATION

The system was studied in detail by analyzing the effect of changing some of its parameters and sources of error. The factors studied were:

1. Length of antenna.
2. Size of wire.
3. Angle between the antennas.
4. Height of antenna.
5. Effect of obstructions.
6. Effect of unequal currents.
7. Effect of polarization.
8. Effect of sky wave.
9. Effect of weather.

Figure 5 shows that not much is gained by making the antenna longer than 100 ft. Although measurements showed that sharper courses would be obtained with longer antennas, the course with the 100-ft antenna is only ± ½° wide which is sufficiently sharp.

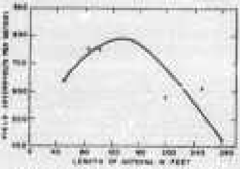


FIGURE 5. Field strength versus antenna length. W-110B wire on ground.

The current distribution was measured to see if standing waves were appreciably reduced when using long antennas. The current in a ground antenna as long as 230 ft was a standing wave having an attenuation of only 25 per cent per wavelength and a velocity of propagation of 0.8 the velocity of light.

No appreciable variation of the velocity of propagation was noted when No. 18 enamel-covered wire, Army wire W-110B, No. 14 stranded insulated wire or No. 14 solid copper rubber-covered wire was used. Although the desirability of using thinner wire to obtain a higher velocity of propagation was evident, experiments with such wire showed it to be impractical for field use.

Effect of Angle Between Antennas

The 90° angle between the antennas of the British system is not essential. A system using a 60° angle was tried and the course obtained was fully equivalent to one using the 90° antennas except that there apparently was a slight amount of coupling between the two antennas in the 60° position so that the field from the energized antenna was diminished about 3 per cent along the direction of the course. The only advantage in using a 90° angle is the greater ease of accurately laying out this angle.

The degree of interaction of the two antennas is shown in Figure 6. Here one antenna was energized and its field measured at a point on a line making an angle of 45° thereto, while the unenergized antenna was swung from a position parallel to one perpendicular to the energized antenna. The maximum variation of the field under these conditions was about 8 per cent. The field strength was 157 (in relative units) when the unenergized antenna was entirely removed. When it was placed perpendicular to the energized antenna and raised from the ground to a height of 8 ft, the field strength varied only from 162 to 164. It may be concluded from these measurements, and from the many 360° surveys of the courses which were made, that a deleterious interaction of the antennas will not occur under any conditions likely to be encountered.

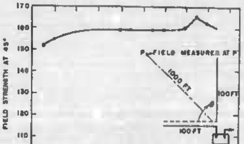


FIGURE 6. Field strength of energized antenna as function of angle of unenergized antenna.

Effect of Antenna Height

Measurements indicated that raising the antenna from the ground to a height of 2 ft increased the field strength in the direction of maximum radiation only about 25 per cent, and that a height of 6 in. produced only a 10 per cent gain in field strength over the case of the

antenna lying on the ground. Therefore it was decided to adopt the simple practice of stretching the antenna along the ground unsupported.

Effect of Obstructions

For this type of beacon obstructions near the transmitter appear to have very little effect on site errors. Such obstructions as a full-scale model airplane 25 ft from one of the antennas did not appreciably affect the course. An automobile placed within 5 ft of one of the antennas also had no measurable effect. A small building having electrical wiring about 20 ft from the apex of the antennas, small trees, a wooden tower 20 ft tall, and another wooden tower 40 ft tall, close to the antennas produced no noticeable deviations of the course.

Obstructions near the receiver site were found to be quite important. Bends in the course resulting in errors as high as 4° were noted in the neighborhood of overhead power lines, and multiple courses were also noted in their neighborhood. In one case a course was traversed by a power line at an acute angle at a distance of 400 yd along the course. There was also at this point on the course a building 150 ft long and 40 ft high having electrical wiring. The course was bent about 4° in the vicinity of the power line. However, the course was found to resume approximately its correct direction about 100 yd beyond this line. The effect of these obstructions was undoubtedly increased by their location on high ground.

There appeared to be some correlation between hills and deviations of the course, but it was inconclusive because of the invariable presence of other site factors. The course was found to be straight through woods. A barbed wire fence across the course at an angle of about 90° did not appear to bend it measurably.

Effect of Unequal Currents

The course of the British beacon lies along the bisector of the angle between the antennas when the antenna currents are equal. When the current in one antenna is greater than the other, the course is deflected toward the other antenna. This effect is utilized in directing the course. Capacitors C_1 and C_2 in series with the antennas, shown in Figure 7, vary the currents in the antennas and thus determine the direction of the course. These capacitors (maximum capacity 100 $\mu\mu f$) are varied differentially by a single control knob, which when turned to the right steers the course to the right, and when

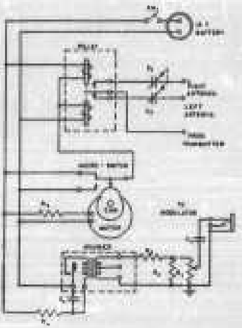


Figure 7. Modulator-keyer unit used with British system.

turned to the left steers the course to the left. The course may be steered $\pm 20°$. The course may also be steered by potentiometers placed in series with the antennas at their sending ends. It is considered more advantageous, however, to use capacitors because of the ease of attaining smooth operation, the avoidance of loss in the potentiometers, and the greater durability of condensers.

Effect of Polarization

The ground antennas in addition to radiating the desired vertically polarized field also radiate a horizontally polarized field. At a distance of 200 yd, at an angle of 45°, with the antenna

3 ft above the ground, the vertically polarized field was 3.5 mv and the horizontally polarized field was 0.47 mv. The horizontally polarized field is capable of producing an error of 2° at 100 yd and 1° at 200 yd. The polarization error at 300 yd is ½° and at greater distances it becomes negligible. These errors represent the maximum deviations which can be obtained with the SCR-536 receiver at heights of about 2 ft above the ground and were determined by locating the apparent course with the receiver held horizontally and at right angles to the course, and then turning the receiver 180° in the horizontal plane and relocating the course. The difference between these two course determinations is called the horizontal polarization error.

Sky-Wave Effect

The course was tested at night and no sky-wave difficulties were noted. The critical frequency of the F layer at Washington at the time was lower than 5.5 mc. Therefore, sky-wave propagation at this frequency could occur only by means of sporadic E-layer clouds. The radiation of the ground antennas in a vertically upward direction is so small that it seems very unlikely that an appreciable signal can ever be received via the ionosphere. The course showed the same accuracy at night as during the day. A loop direction finder placed on course at a distance of 2,500 yd showed no evidence of polarization error. There was also no evidence of fading at this distance, using a meter in the output circuit as an indicator.

Effect of Weather

This British type of beacon was tested under various weather conditions including heavy rainfall and while the ground was covered with a light snow. Such weather conditions did not appreciably shift the course. The course shift from a dry day to a succeeding rainy day measured at a distance of 700 yd was only ¼°, which is within the limits of experimental error.

Modified British System

The equipment for this system differs from that of the British system chiefly in the antennas. The antennas for the modified British system consist of a ground antenna at 90° to the desired course, and an antenna for radiating vertically polarized waves of the proper phase with respect to the field from the first antenna. The second antenna may be the vertical rod antenna normally used with the SCR-284 transmitter. Three antenna systems for the modified British beacon are shown in Figure 8. In Figure 8A the ground antenna at right angles to the course is a dipole type of Beverage

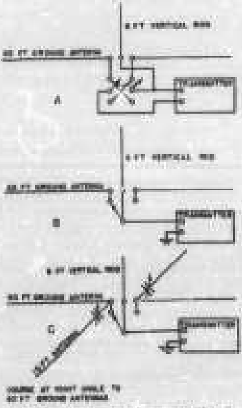


FIGURE 8. Arrangements of antennas with modified British system.

antenna. The antenna system in Figure 8B consists of a pair of single-ended Beverage-type antennas arranged in a straight line at right angles to the course. The course is obtained by

reversing the current in the 90° antennas and thus switching the radiation pattern.

The modulator and keyer unit for the modified British beacon is similar to that of the British beacon. A circuit diagram is shown in Figure 9.

The course of the modified British beacon can be shifted by suitable means. One such means which has been thoroughly tested is shown in Figure 8C and consists of a pair of antennas about 15 ft long arranged along the ground at right angles to the main ground antennas. These antennas are fed through a pair of uncontrolled capacitors which vary the current in these antennas. By means of these capacitors and a reversing switch (SW_1 in Figure 9) the course may be shifted about 7° to the right or left.

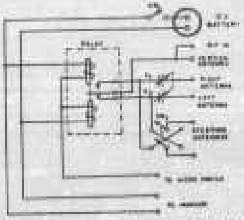


FIGURE 9. Connections of antennas and relays with modified British system. [illegible] Figure 7.

Purpose of Modified British System

The original purpose of the modified British beacon was to eliminate key clicks by performing the required antenna switching in those antennas which have a null in the direction of the course, and thus prevent the switching from affecting the field in this direction.

Another purpose was to eliminate those errors which are caused by factors that may cause the ground antennas of the British beacon to radiate fields of different intensities or different directional characteristics. Such factors are variations in height of the antennas, length, ground over which the antennas are stretched, and current distributions of the antennas. The independence of the modified system of these factors affecting the intensity of the radiated field arises from the fact that the course in this system is determined only by the location of the null of the radiation pattern and not by the absolute magnitude of the field or the shape of the radiation pattern.

Factors Affecting Operation. The modified British system was studied by analyzing the several parameters and factors which might cause errors.

Ground antennas of various lengths were studied, both of the dipole and single-ended types. These studies showed that a dipole 120 ft long, and a single-ended antenna 60 ft long had a sharp null and a high ratio of vertical to horizontal polarization. Figure 10 shows the radiation patterns of several types of antennas in the neighborhood of the nulls and exhibits the superiority of the 120-ft dipole at 5.5 mc. Similar tests showed that 60-ft and 100-ft single-ended antennas had satisfactory radiation patterns. These curves, of course, are applicable only to antennas utilizing the type of

FIGURE 10. Radiation pattern in vicinity of nulls of several antenna types.

wire and arranged at the heights above ground used in these tests.

The effect of the size of wire and the height above ground has already been discussed, and the same findings which apply to a single-ended antenna also apply to a dipole type of ground antenna.

The effects of obstructions both at the trans-

mitter and receiver sites are the same as for the unmodified British system.

Tests made to determine the effect of unequal antenna currents on the position of the course showed that even a ratio as great as four to one between the currents in the main ground antennas had no effect on the location of the course. This is very important because unequal antenna currents always occur.

The effect of horizontal polarization in causing errors was found to be the same as in the British system.

The height of the vertical rod antenna and the current in this antenna determine the range of the beacon and the sharpness of the course. With ground antennas 60 ft long a 9-ft vertical antenna gives approximately the same range and sharpness of course at a frequency of 6.5 mc as the 45° antenna of the British beacon.

11.3 DESIGN OF SWITCHING RELAY

It was early recognized that one of the chief problems in perfecting the British beacon was that of eliminating key clicks. One solution was the use of the modified British system, the other was the design of a special relay.

Many variations of relays were tested. The first attack on the problem was to study the cause of the clicks. It was found that the clicks resulted from the fact that during the time of commutation the r-f current in the antennas was reduced to a low value and that the intensity of the clicks was dependent on the time during which the current remained low. The designs were, therefore, directed toward a relay in which the actual time of commutation was reduced to a minimum. In commercial practice this result has been achieved by the use of very large magnets operating small moving parts. Relays weighing as much as 20 lb have been used for this purpose. In the present case such relays were not practicable. It was also found undesirable to develop a relay which would reduce the current in one antenna gradually before increasing the current in the other antenna. The effect of such a shift was to reduce the accuracy with which the course could be detected. To reduce the time of commutation it was realized that the moving contact would have to be made as light as possible and that commutation should take place only after this moving contact had reached a substantial velocity. It was also found important that the inertia of the magnetic armature carrying the moving contact should cause as little delay on the action as possible.

The first relay tried, while suitable for the voice-modulation type of British beacon, was found entirely unsuitable when tone modulation and A-N keying were applied. An antenna change-over relay having a 1/300-second time of throw was also found entirely inadequate. Snap-action switches such as a microswitch produced pronounced clicks. Two switching arrangements having make-before-break action gave no improvement.

A double-contact switch making contact first through a resistor and then making contact directly was tested. The resistors were varied from 0 to 2,000 ohms and the least click appeared to occur when the resistors were entirely out of the circuit. This scheme appeared to have no promise.

An Allied Control Company Type AK relay modified similarly to one used by the Naval Research Laboratory for the same type of beacon and loaned for study was tested and found to have a change-over time of approximately 1/2,000 second. This relay would probably give very little key click. The result was obtained by an excessive current in the magnets causing them to become extremely hot. Since this type of relay was no longer manufactured, no work was done to incorporate it in the experimental models finally submitted.

The relay shown in Figure 11 was built along the principles of having a light contact reaching a substantial velocity before commutation. It could be adjusted to have a change-over time of about 1/4,000 second. To avoid chatter it was necessary to dampen the vibration of the movable contacts by a packing of Airfoam sponge rubber. This relay was connected across a 12-volt battery which was switched from coil to coil by a microswitch. The current taken was 1.5 amperes. The relay armature and movable contacts would normally occupy an intermediate rest position where they do not connect to either fixed contact during the throw of the microswitch, because the microswitch takes a comparatively long time to move from

one of its contacts to the other and the coils of the relay are unenergized for a sufficient time to allow the relay movable contacts to return to this center position. This condition was at first remedied by adjusting the magnetic circuit so

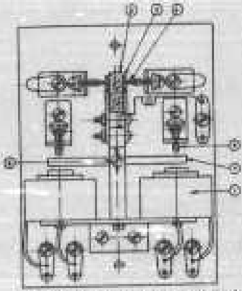


FIGURE 11. Details of keying relay as developed for beacon: (1) coil, (2) Airfoam rubber pad, (3) fixed contacts, (4) movable contact, (5) armature stop screw, (6) adjustable bearing, (7) armature.

that there was sufficient residual magnetism to hold the armature in one position until the coil in the other position was energized. Later the relay circuit was modified so that both coils are energized during the change-over of the microswitch and hence the armature is positively held in a given position until the coil in that position is short-circuited.

The relay is adjusted as follows. The fixed contacts are screwed in about 0.003 in. beyond the point where they just touch the movable contacts. During operation the movable contacts acquire such a high velocity just before making contact on the other side that the closed movable contact cannot, by virtue of its spring, remain closed. The best adjustment of the relay appears to be one in which the change-over time is about 0.00025 second. When so adjusted both antennas are simultaneously connected to the transmitter for the very brief time of about 0.0001 second, but this appears to have no bad effects. Armature stop screws maintain a sufficient gap between the relay armature and pole pieces to prevent the armature from locking in one position.

The relay can easily be set to have a change-over time of 0.0003 second. It showed no deterioration of performance after 24 hours of continuous operation. These results require a positioning of the fixed contacts with a tolerance of the order of 1/1,000 in. The models delivered to the Signal Corps were laboratory models and might not be able to maintain this degree of tolerance under hard field use. The Navy Department advised, however, that such a degree of tolerance can be maintained satisfactorily in the field. It is believed, therefore, that with suitable mechanical improvement in the design there will be no difficulty in having reliable relays for field operation.

11.8 SETTING UP THE ANTENNAS IN THE FIELD

Several methods of installing the antennas in the desired directions were developed and mechanical aids were delivered with the equipment. These included a magnetic compass with a pair of sights and aligning bars. The antennas can be aligned by placing them approximately in the correct directions and then by adjusting the currents so that "steering" occurs. These methods are described in greater detail in the contractor's report on the project.

11.9 EQUIPMENT DELIVERED

Three sets of the final model of the equipment for the British beacon together with an antenna aligning device were delivered August 5, 1944. Each set of equipment consisted of an instruction book and a carrying case containing two reels of antenna wire, two antenna ground stakes, cables, and a modulator-keyer unit.

The modulator-keyer unit was housed in a waterproof box 10x7x9 in. The 1,000-cycle tone used for modulating the transmitter was generated by a General Radio Type 572-B hummer. The two coils of the keying relay were

connected across a 12-volt battery. A cam-operated microswitch short-circuited the coils alternately. The cam was cut to produce interlocked A and N characters and was rotated by a Haydon timing motor. The speed of the motor was adjusted to produce a keying rate of 64 characters a minute. The motor is capable of keying up to 128 characters a minute by a suitable adjustment of a resistor. A pair of 100-$\mu\mu$f straight-line frequency capacitors were used for setting the course within 20° of either side of the bisector of the angle between the antennas. The capacitors were coupled together so that they were rotatable by a single knob and were arranged so that they varied differentially but had equal capacitances at a midposition.

The unit was powered by the 12-volt battery used for the transmitter. Its power consumption was 3.7 watts. It was capable of producing 100 per cent modulation of the transmitter at the maximum power output of the transmitter. The weight of the complete unit in a heavy steel box was 18 lb. This weight could be reduced to 9 lb by the substitution of an aluminum box and a reduction in size of the unit.

11.10 CONCLUSION

Of the three types of beacons studied experimentally the British and modified British beacons were found superior to the crossed-loop beacon. The experimental models of the British and modified British beacons gave substantially equal results. The modified British system is less subject to certain possible sources of error, but it is slightly more complicated if a steering adjustment is required. From a designer's point of view the modified British system has also the advantage that the course can be readily broadened or sharpened at will by altering the ratio of the currents in the vertical and horizontal antennas. As the course is broadened the signal intensity on the course is increased and vice versa.

A comparison of the three types of beacons is presented in Table 1.

TABLE 1. Comparison of beacon systems.

	Crossed loops at 0° and 90° to required direction	British beacon	Modified British beacon
Field strength at 1.5 miles	42 μv/m	60 to 100 μv/m	60 to 100 μv/m
Width of course for ±1 db	±2° with best installation	±1°	±1°
Tuning procedure	Least simple	Simple	Simple
Steering of course	Readily done	Readily done	Readily done
Key-click elimination	No special relay required	Requires very rapid, well-adjusted relay	No special relay required
Unbalance between A and N currents	Causes no error	Current Ratio / Error: [illegible] 2°; [illegible] 9°; [illegible] 25°	Causes no error
Time required to install and tune	More than 10 minutes	Less than 10 minutes	Less than 10 minutes
Weight exclusive of transmitter	Approx. 40 lb	Approx. 30 lb	Approx. 30 lb
Polarization error	Not measured	±2° at 100 yd ±1° at 200 yd 0° at 400 yd	±2° at 100 yd ±1° at 200 yd 0° at 400 yd
Variation of ground under the 2 antennas		Shift possible	0°
Unequal length of antennas		Lengthening 220-ft antenna 16 ft caused shift of 3° at 100 yd, 5° at 200 yd.	Lengthening 90-ft antenna 10 ft produced no measurable course shift

Chapter 12

U-H-F DIRECTION-FINDING ANTENNA STUDY

Development of a direction-finding system' covering the range 140 to 600 mc, providing instantaneous bearing indications for vertically polarized signals. Two wave collectors utilize a common receiver and indicator. One antenna consists of an Adcock system with output fed into a capacitive goniometer; the other antenna (for 300 to 600 mc) is a rotating element in front of a reflector, the position of the antenna being synchronized with the CRO indicator.

12.1 INTRODUCTION

The object of this project[a] was, briefly, to develop a d-f system operating in the u-h-f region of 140 to 600 mc. It was hoped that much of the experience gained and the means developed in previous development programs on d-f systems for lower frequencies (1.5 to 30 mc) could be drawn upon in this project. It was found, however, that while the experience was useful, the methods employed in the lower-frequency systems so usefully could not be effective in the u-h-f region.

12.2 PROBLEM DEVELOPMENT

In the systems developed for the 1.5- to 30-mc region, aperiodic thermionic (cathode follower) coupling between the high impedance of the antennas and the low-impedance lines connecting the antennas to the receiver was quite effective in making it possible to space the receiver at some distance from the antenna, and to provide an impedance match between antenna and line. An attempt to use this method in the higher-frequency region failed for the simple reason that tubes available at the time provided no more energy transfer when the tubes were operating normally than when they were cold. The major contribution to transfer existed in the capacitances within the tubes.

It was found also that an inductive goniometer had to be abandoned because the transfer through it was largely capacitive and because of its low impedance.

An electronic goniometer depended upon obtaining identical transfer characteristics through four separate tubes at all points of a modulation cycle. The difficulty of matching tubes made it impossible to obtain equality of transfer with modulation or to obtain adequate transfer of energy over the wide frequency range contemplated. This system had to be abandoned. Since the inductive goniometer behaved better as a capacitive than as an inductive instrument, further work was concentrated on the development of a truly capacitive goniometer with the result that adequate transfer was obtained. The final model of the direction finder employed such a unit.

Using the design principle which had previously proved adequate in the frequency range 1.5 to 30 mc, a ground plane carrying four monopole antennas, acting in pairs to give crossed figure-eight diagrams, was constructed. Since the thermionic coupling means were proved to be unsatisfactory the antennas were terminated resistively.

The receiver research for this project passed through three stages. The preliminary receiver was constructed having one r-f stage, an oscillator and mixer each tuned by means of coaxial lines the movable elements of which were ganged to a single control. The r-f input of this receiver was applied through a 50-ohm coaxial transmission line.

The first modification was alteration of the input circuit to obtain balanced input. The second and final modification consisted of a complete mechanical redesign to avoid the necessity for having the cumbersome tuning method of the previous models.

12.3 SYSTEM EXPERIMENTS

The first experiments with the complete d-f system were conducted using a capacitive goniometer mounted on the Type A indicator[b]

[a] Project C-80, Contract No. OEMsr-961, Federal Telephone and Radio Corporation

[b] The Type A indicator utilizes a cathode-ray tube and circular trace. The trace is obtained by mechanically rotating magnetic deflection coils about the neck of the tube. The rectified received signal is fed into the coils to change the circular trace to the typical propeller-shaped direction pattern.

in place of the normally used low-frequency goniometer. The antenna output was connected to two balanced transmission lines, one for each antenna pair, and applied to the two sets of stator plates of the capacitive goniometer. The first system tested was composed of the most satisfactory elements determined from the preliminary research. The monopole antennas were resistively terminated. Use of two 40-ft balanced transmission lines enabled the collector system to be placed at a distance from the receiving and indicating equipment. It was immediately determined that very poor nulls were secured, that the nulls were not reciprocal and that the overall sensitivity of the system was very poor. A modification program was instituted leading to the following changes:

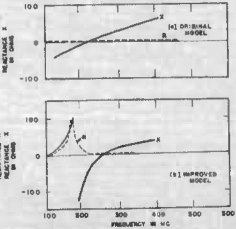


FIGURE 1. Characteristics of capacitive goniometer.

The first capacitive goniometer used inductive means for coupling the rotor plates to the receiver input. This output transformer gave very poor transfer and immediate steps were taken to increase the efficiency. One goniometer was constructed in which the inductive output device was replaced by slip rings. A considerable gain in transfer was apparent but due to the use of a continuously rotated goniometer, the slip rings required frequent maintenance. A capacitive-output coupling system was then constructed which gave reasonably good transfer characteristics. The characteristics of the capacitive goniometer are shown in Figures 1 and 2.

FIGURE 2. Transfer characteristics of goniometer.

One of the principal reasons for poor nulls and for nonreciprocal bearings was the fact that the transmission lines connecting the antennas of one pair were not properly shielded

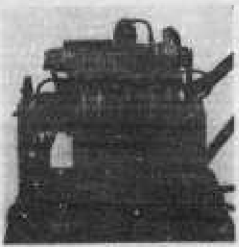
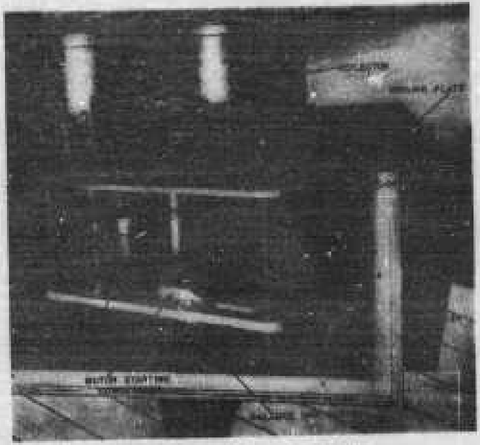


FIGURE 3. Direction-finder receiver in which tuning is accomplished by rotating a drum which supports entire r-f and converter section, varying effective length of three coaxial lines and one quarter-wave open wire line (local oscillator). Four-stage 15-mc i-f amplifier with gain of 25,000 and band width of 1 mc follows the converter.

and that there was direct pickup on these lines. It was found necessary to shield very thoroughly the transmission lines themselves, to provide additional shielding at the crossover

point, and to enclose the entire transmission line system within an additional shield.

After the shielding means had been employed poor nulls were still observed over a considerable portion of the frequency range and large errors were introduced because of unbalance in the 40-ft transmission lines between the collector system and the goniometer and because of the differences in electrical length of these two lines. Therefore the capacitive goniometer was moved into close proximity with the antenna system. A further improvement was effected when the output of the goniometer was fed directly into a "balance box" transforming the balanced output of the goniometer to match the unbalanced coaxial transmission line. This modification not only enabled the distance between the antenna and the receiver to be increased but in addition eliminated a great many of the poor nulls which had previously been observed.



[illegible]

At the same time it was possible to begin tests with an improved model of the receiver (Figure 3) having square cross-section transmission lines as the tuning elements coiled on a drum which was rotated by the dial mechanism. This receiver used lighthouse tubes throughout and was more sensitive than previous models.

Some difficulty was encountered because of the use of sliding short circuits as tuning elements of the transmission lines.

To localize any difficulties which might be contributing to errors or to poor operation, an extensive series of tests was instituted on the separate components of the collector system to determine the impedance characteristics of each over the frequency range and, if possible, to discover design criteria. The results obtained showed that the antennas would be extremely

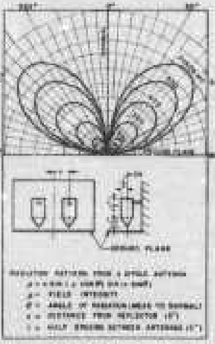


FIGURE 5. Calculated field patterns of two vertical monopoles mounted in front of infinite reflector. Monopoles are fed 180° out of phase with each other. At mean frequency, half-space between antennas was [illegible] in., and spacing to reflector, [illegible] in. Ratio of field intensity at 600 mc to that at 300 mc is almost [illegible] for equal fields set up by antenna.

difficult to match to a transmission line and indicated why the capacitive goniometer ceases to function at about 300 mc and in general show the difficulties which were encountered in an attempt to make a monopole system of this type operate over such a wide frequency range without drastic changes in design.

12.4 FINAL DESIGNS

For several months studies had been in progress on a collector system constituted by two oppositely connected dipoles spaced from each other and in front of a reflecting plane surface. To obtain automatic instantaneous indication from a system of this type, a collector was constructed as illustrated in Figure 4. This rotating collector was driven by a large induction motor and the instantaneous position of the collector was repeated through a selsyn system so as to be shown on a CRO screen. The calculated directional pattern of the collector, the measured pattern and the resulting indication are shown in Figures 5 and 6. The system operated with satisfactory results between 300

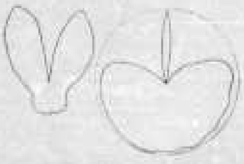


FIGURE 6. Measured field pattern (left) and resulting indicator pattern of monopole-reflector system.

and 600 mc, thereby supplementing the performance which had been obtained using the fixed monopole system and the capacitive goniometer.

As a final step in the development, the low-frequency system (140 to 300 mc), consisting of the five monopole antennas and the capacitive goniometer, and a high-frequency system (300 to 600 mc), consisting of the rotating antenna, were incorporated for use with a single control unit consisting of the receiver, an indicator and the necessary power supplies.

PERFORMANCE

Figure 7 shows that the sense performance of the 140- to 300-mc monopole system is not adequate. A considerable amount of redesign and further development would be necessary to obtain results which would permit a production-type system to be built. Figure 8 shows the directional accuracy of the monopole antenna collector system with the capacitive goniometer in the frequency range 140 to 300 mc. This performance could also be considerably improved.

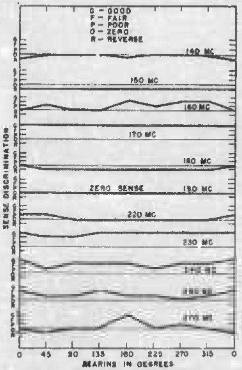


Figure 7. Sense operation characteristics of 140- to 300-mc Adcock.

The rotating antenna system is not subject to the same type of errors as the monopole system. The accuracy is indicated as $\pm 3°$ in all tests made. No sense ambiguity is possible with this type of collector system. Between 300 and 600 mc, nulls are always sharp and in every way the operation of this system is much more satisfactory than that of the fixed-monopole system.

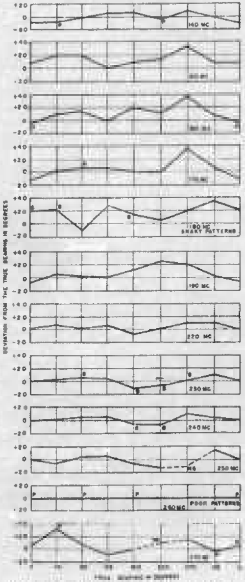


Figure 8. Bearing accuracy of 140- to 300-mc Adcock antenna.

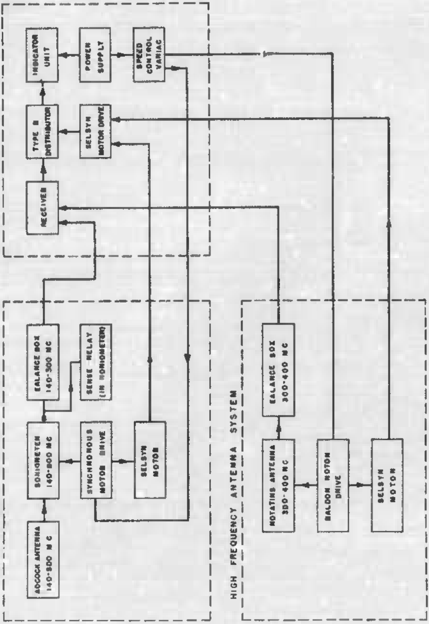


FIGURE 9. Block diagram of 140- to 600-mc direction-finder system.

12.6 ELECTRICAL CIRCUIT THEORY

The entire system as finally developed consists of three major units: Band I (140 to 300 mc) wave collector and goniometer; Band II (300 to 600 mc) wave collector; receiving and indicating unit for remote operation. (See Figure 9.)

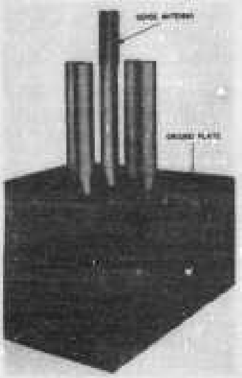


Figure 10. Low-frequency (140- to 300-mc) antenna system.

Band I Wave Collector

As shown in Figure 10 the 140- to 300-mc Adcock wave collector consists of five vertical monopoles mounted on insulators over a copper ground plate. Directly below the plate and mounted in a wooden protective box are the capacitive goniometer, the driving motor, and the selsyn generator. The entire system block

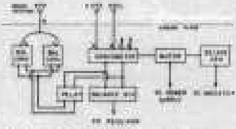


Figure 11. Elementary schematic of low-frequency portion of direction-finder system.

diagram is shown in Figure 11 where only one Adcock pair is indicated for the sake of simplicity. The polar diagram of this array is a figure eight (Figure 12).

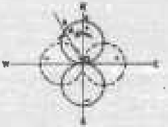


Figure 12. Figure eight pattern of Adcock antenna.

This pattern follows a cosine function. If the antenna system were rotated by hand only one pair of antennas would be required, the position of the nulls indicating the direction of the received signal. For instantaneous indication the capacitive goniometer scans the output of two pairs of Adcocks (four antennas).

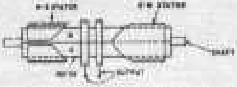


Figure 13. Elements of capacitive goniometer.

Goniometer

The rotor of the goniometer consists of two semicircular plates, A and B, in Figure 13, insulated from each other. The two pairs of stators are identical except that one is oriented 90° with respect to the other. One output ring is connected solidly to rotor A, while the other is connected to rotor B. These are rotated inside two fixed rings to provide capacitive coupling to the rotor output. (See Figure 14.)

The stator plates are so shaped that the capacitive coupling between rotor and individual stators varies as a cosine function with rotation. For example, assume both pairs of antennas connected to both stators and the signal being received in the N-S direction. The signal will be in the null of the E-W antennas so that no voltage appears across the E-W stator to be picked up by the rotor. For the N-S stator, as the rotor is turned slowly, the output will vary from a maximum when the plates A and B are parallel to the stator to a minimum of zero when the rotors are at right angles to the stators. Thus two nulls are produced 180° apart.

Similarly, if the signal is in the direction of the E-W antenna, two nulls will again be produced 180° apart, except that they will be 90° away from the previous ones because of the positioning of the E-W stator. In this manner the goniometer will indicate bearings of signals in line with the antennas.

For the case where the signal direction is not in line with either array, assume the signal is received along the line o-b (Figure 12). This means that there will be o-a volts delivered to stator E-W, and o-b volts delivered to stator N-S. Therefore, across stator N-S there will be a voltage

$$e \cos \theta,$$

where e = voltage (o-c), and across stator E-W, there will be a voltage

$$e \sin \theta.$$

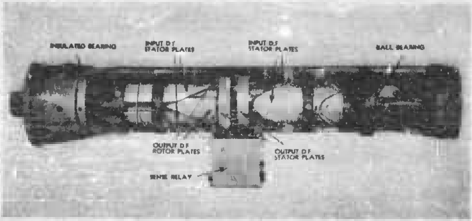


Figure 14. Photograph of goniometer.

For any rotor position, there will be a voltage coupled from stator E-W to the rotor proportional to $e \sin \theta$ and equal to

$$K\,[e \sin \theta].$$

Similarly, the voltage coupled from stator N-S will be equal to

$$K[e \cos \theta].$$

If the rotor is lined up for maximum coupling to stator E-W, and then rotated through an angle β, the voltage across it will be

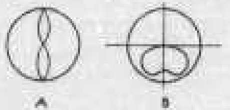
$$K[e \sin \theta][\cos \beta].$$

The voltage coupled into the rotor from stator N-S will vary from zero to maximum as β is increased and the resultant voltage will be

$$K[e \cos \theta][\sin \beta].$$

There will be some position for a rotation of β degrees where

$$K\ [e \cos \theta]\ [\sin \beta] = -K\ [e \sin \theta]\ [\cos \beta].$$



[illegible]

At this point, the voltages will cancel in the rotor for zero output in a null position. Solving the above equation, it will be found that

$$\theta = -\beta$$

indicating that the position of the rotor indicates the actual bearing. This would produce a double null pattern as shown in Figure 15A. To establish sense, the output of the sense antenna must be fed in phase to the goniometer output to produce a cardioid pattern. Since the Adcock monopoles are cross-connected, an analysis of the voltage vectors will show that the sense antenna output is 90° out of phase with the Adcock antenna output. To shift it 90°, the output of the sense antenna is fed through two unequal transmission lines, through unbalance-to-unbalance converters, and mixed through a relay with the goniometer output. The line

[illegible] 17. Block diagram of high-frequency antenna system.

lengths are so proportioned as to produce a 90° phase shift over the band. When the sense antenna is connected in the circuit the sense pattern would theoretically appear as shown in Figure 15B where the pattern indicates the direction of the bearing. Because of difficulties with balance in two coaxial lines, the goniometer output is fed into a balance-to-unbalance converter and then via a single coaxial line to the receiver.

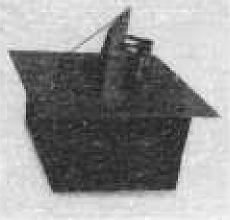


FIGURE 16. High-frequency (300- to 600-mc) rotating antenna.

FIGURE 18. [illegible] pattern from crossed high-frequency [illegible]

The goniometer is rotated by means of a motor which also turns a selsyn generator. This selsyn generator is used to drive a selsyn motor in the indicating unit.

BAND II WAVE COLLECTOR

As shown in Figure 16, the 300- to 600-mc collector consists of a pair of vertical monopoles

in front of a reflector and rotated over a ground screen. A block diagram of the system is shown in Figure 17.

The entire system is mounted in such a manner that the cylindrical-shaped balance box serves as a shaft for rotation. The output is taken off via a fixed line about which the balance box rotates so that no rubbing contacts are used. The monopoles are cross-connected at the balance box so that the antenna patterns are approximately as shown in Figure 18.

Since a sharp null is produced in the direction of the received signal, the system is unidirectional and requires no sense, as in the Band I collector. As the collector is rotated, the operator will first find the signal over a rather broad lobe, pass through a sharp null, continue over another broad lobe of reception and then pass through approximately 180° of null. The collector is driven by a variable-speed motor which also drives a selsyn generator for synchronization with the indicator.

RECEIVING UNIT

The receiving unit (Figure 19) consists of a 140- to 600-mc tuned line receiver, a d-c amplifier, and switching circuits for operating Band I and Band II collectors. Tuning the receiver is accomplished by varying the length of a circular transmission line by means of shorting bars. Receiver input is single-ended and is fixed at 90 ohms. The i-f channel is straightforward and has a bandwidth of 1,000 kc for passage of pulses.

Motor switching circuits are interlocked so that only one system can be operated at a time. In operation only the r-f cable need be changed for a band change.

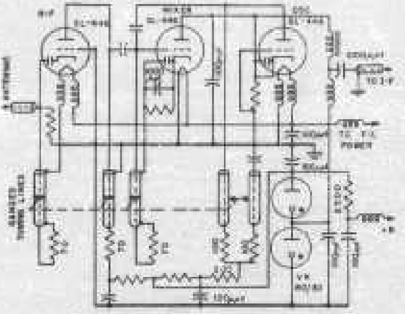


FIGURE 19. Essential elements of receiver circuit diagram.

INDICATOR UNIT

The Type B indicator (Figure 20) with two selsyn motors for driving, and speed control for antenna systems are mounted on the power-supply chassis. The Type B indicator consists

of a strip of alternate thin laminations of copper and polystyrene. The projecting ends of the laminations are ground to a flat surface and a uniform resistance strip is compressed on one side. This produces a commutator with a large number of equal resistance steps between bars. The strip is rotated by a pair of selsyn motors to produce the voltage needed to generate a circular trace in the cathode-ray oscilloscope.

If a current is sent through the strip a sinusoidal voltage will be generated across a pair of brushes mounted along a line perpendicular to

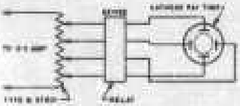


Figure 20. Elements of Type B [illegible]

the rotational axis and equidistant from it. By mounting ahother pair of brushes at right angles to the first pair, two sinusoidal voltages are obtained with 90° phase difference. These voltages applied to the deflecting plates of the CRO tube produce a circular trace when the spot moves at constant velocity.

By supplying the Type B strip current from the plate of a d-c amplifier following the receiver detector, the receiver output can be 1, ade to vary the shape of the circle for an indication.

When the receiver output is zero at the null (0°) the plate current in the d-c amplifier will be maximum and the spot will be at the outside of the circle. As the goniometer scans from 0° to 90°, the receiver output will increase to maximum, biasing the d-c amplifier until cutoff is reached, and no voltage will appear across the strip. Thus the spot will approach the center, rapidly at first because of the sharpness of null and then more gradually.

For sense operation the same principles apply except that the cardioid pattern resultant produces a pattern with one broad null.

To place the cardioid pointing in the same direction as the d-f pattern, it is necessary to turn the pattern by 90° on the cathode-ray tube. This is done b, means of a four-pole double-throw relay which switches each brush to the adjacent cathode-ray tube deflection plate. Positioning of the circle is effected by magnetic deflection coils placed about the neck of the cathode-ray tube and operated from the low-voltage supply.

Circle diameter is varied by cathode bias control of the d-c amplifier. Speed control is incorporated into the wave collector motor since it is necessary to bring the selsyn motors up to speed gradually.

Chapter 13

LOCATING TANKS BY RADIO

Problem of locating the position of friendly tanks with respect to a fixed station to an accuracy of ± 50 yd in 5 miles using existing Signal Corps tank equipment by an audio-phase-measurement method. Investigation of the characteristics of existing tank equipment indicated that inherent phase instability would make impossible location of tanks to the required degree of precision.[1,2]

13.1 INTRODUCTION

THE BASIC IDEA involved in these two projects[a] was to place a constant audio tone on the carrier of a standard communication transmitter at a locator station. This signal would be received by the tank and the tone would be retransmitted by the tank on another radio frequency. Assuming constant time delay, or phase shift, through the transmission and reception networks, the measured phase shift in the audio tone as measured at the locator station would be a measure of the distance between the tank and the locator station. The location of the tank or group of tanks would be accomplished by a triangulation process.

One requirement established was that existing equipment be employed in these projects. Therefore, although the method for locating tanks by radio was considered basically workable, whether the scheme would be successful would depend entirely upon the following two major factors:

1. The accuracy with which the phase measurement could be made.
2. The stability of the phase shift through the tank equipment under normal operating conditions.

Tests, therefore, were made by the two contractors on the phase stability of two existing pieces of radio equipment, the SCR-506 in the 2- to 4½-mc region and the SCR-508 in the 20- to 30-mc region.

[a] Project C-60, Contract No. OEMsr-787, Bell Telephone Laboratories; and Project C-61, Contract No. OEMsr-737, General Electric Co.

13.2 TEST RESULTS

13.2.1 Tests on SCR-506

To measure the distance of the tank within ±50 yd at 5 miles requires an accuracy of 0.57 per cent. Using an audio frequency of 2,000 cycles per second would result in a phase shift of 38.7° for a 5-mile spacing between tank and fixed station. To measure this phase shift to an accuracy of 0.57 per cent would require that measurement to 0.22° would be necessary.

Measurements on the SCR-506 (Project C-61) were accurate to about ±0.25°. It was found that the slope of the tuning curve of this receiver was about 1° per kc off tune. Using the beat-frequency method, this error might be held to 0.05°. Even when the local oscillator was adjusted by the zero beat method, a change of phase shift of 0.07° occurred per degree centigrade rise in ambient temperature. The average slope of the curve of phase shift versus percentage modulation was about 0.12° for a 1 per cent change in modulation. With the automatic volume control disconnected (manual gain control condition) severe phase shifts with changes in signal level occurred. In the a-v-c condition, no measurable phase shift occurred with a signal level change of 10 to 1. A signal level of at least 1,000 μv would be required for reliable readings. In the operating region, the slope of the volume control setting curve showed a phase shift of approximately 0.12° per degree rotation of the volume-control knob.

In light of these measurements, it was decided that the instability in phase shift through the receiver alone under normal service conditions would make the audio phase shift method of measuring distance impractical.

13.2.2 Tests on SCR-508

Using the measurement of time as a concept of the measurement of distance, phase shift

would have to be measured within time intervals of 0.306 μsec to accomplish the accuracy of 0.57 per cent required. Direction would have to be measured within 19.5 minutes.

It was found that the inherent variations of phase shift in the SCR-508 (Project C-60), if uncontrolled and uncalibrated in the mobile tank at the time of measurement, would prohibit measurements within ±8 μsec. For example, variations in temperature between −20 C and +50 C together with changes in humidity would produce oscillator drift as much as 50 kc. This alone makes it impossible to meet an accuracy of ± 5.4 μsec or 0.5 mile in 5 miles. Through inability of the receiver's pushbutton tuner to be reset at the same oscillator frequency by merely selecting the same pushbutton would produce an error of ±2.7 μsec. These figures do not include the inherent differences between tank equipments of the same model numbers.

So far as the SCR-508 was concerned, it was apparent that the a-f phase-shift measurement method of measuring distance could not be more accurate than about 25 per cent, or to within 2,200 yd of 5 miles instead of the required 50 yd.

Modification to Improve Accuracy

Variations in the receiver's pushbutton tuners gave errors in excess of 10° at 10,000 cycles. To offset these errors together with the 50-kc oscillator drift would require a crystal-controlled oscillator in the receiver.

By a technique which called for the transmission of two audio frequencies somewhat greater accuracy could be attained since distance would now be determined by the total measured phase difference between the two frequencies rather than the absolute value of phase at either frequency. Assuming that the phase-shifting networks were individually adjusted for each mobile tank installation and that each receiver had the necessary crystal oscillator modifications, an accuracy of approximately 12 per cent or 1,000 yd in 5 miles would be possible.

Elimination of all audio amplification, using the i-f voltage to drive the transmitter, and by making other changes to the receiver (such as changing the intermediate frequency) might result in a phase-shift time in the mobile unit of approximately 4.0 μsec. The amplitude stability of the SCR-508 equipments will not permit the adjustment of two voltages required for measuring phase by the sum-and-difference method to closer than 0.2 db with the result that an accuracy of measurement of 250 yd in 5 miles is about the limit possible with the modified receiver suggested.

Required Measurement Accuracy. A 1° accuracy when measuring phase will permit apparent errors of 90 yd in 5 miles at a modulation frequency of 5 kc. If the modulation frequency is 15 kc this 1° accuracy of measuring phase shift will permit measurements to within 30 yd at 5 miles. Therefore any phase shift method must have an accuracy of 1° or better, particularly if any latitude is to be left for variations at the mobile tank. Such methods are known but they are not of such nature that they could be used in the field easily. Laboratory methods exist which will provide an accuracy of 0.2°.

13.3.3 Simplified Radar Method

The final report on Project C-60[1] proposes a modified radar method in which the tank carries a repeater made up of a 90-db voltage amplifier and a 50-watt 50-mc power amplifier. The fixed station transmits pulses of 1 μsec duration. With a receiver band width of approximately ±3 mc an accuracy well within the prescribed 50 yd independent of the distance measured is estimated. The tank unit being a repeater requires no tuning or crystal and could be readily adapted to equipment already in the field. Thus it would be much simpler than the proposed a-f phase-shift method.

Chapter 14

U-H-F FRIENDLY AIRCRAFT LOCATOR

A d-f system providing automatic and continuous indication of bearings of signals in the region 100 to 250 mc, with arrangements for remote display of the azimuthal distribution of received signals.[1]

14.1 INTRODUCTION

At the time this project[a] was started radar was in its infancy but it was realized that means for identifying friendly aircraft were needed. It was believed that d-f methods giving the azimuth of the target would be useful, particularly if two or more d-f stations could use triangulation techniques.

Means were developed for taking bearings in a matter of about five seconds with an accuracy of approximately $\pm 3°$ and for transmitting the bearing data over conventional telephone facilities. The system was operable on c-w, i-c-w, and pulse signals. Bearings were taken at nearly maximum signal level rather than at a null, and could be taken on two or more signals at the same time provided the bearings were not too close together in azimuth. There was no ambiguity regarding sense. The visual indicator (CRO) traced a polar diagram of the received signal, and an electrical marker system put markers on the CRO screen at 1° intervals.

14.2 THE OVERALL SYSTEM

Principal components of this direction finder consisted of a rotating directional and nondirectional antenna assembly, a u-h-f receiver having two channels for amplifying the respective antenna signals, line transmitter goniometer units to prepare the signals from the d-f channel of the receiver and signals from the goniometers which indicate antenna orientation for transmission over a telephone line, and a line receiver indicator unit which obtained signals from the line transmitter (directly in the case of the monitor and over the telephone line in case of remote operation) and prepared them for tracing out the necessary patterns on the CRO screens for indicating the bearing. A block diagram of the apparatus is shown in Figure 1.

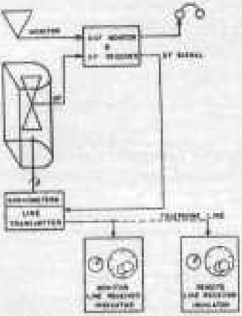


FIGURE 1. Block diagram of [illegible]

14.3 ANTENNAS

The antennas provided (1) a directional lobe of the received signal for bearing purposes and (2) a nondirectional signal for audible monitoring and for a-v-c purposes. The directional antenna consisted of a conical dipole $\lambda/4$ from the origin of a parabolic reflector; as the antenna rotated, a varying signal was induced in the antenna producing a single-lobe pattern with the axis pointing toward the received signal. The antenna rotated at 100 rpm producing rapidly recurrent patterns so that continuous indication of the received signal took place.

[a] Project C-12, Contract No. NDCrc-193, Hazeltine Electronics Corp.

The nondirectional antenna consisted of a single-cone monopole and an artificial ground mounted on top of the parabolic reflector.

Sheathed transmission lines matched to the characteristic impedances of the antennas connected the antennas to the inputs of the receiver. Rotating transformers designed as band-pass filters coupled the transmission lines from the rotating structure to the receiver.

14.4 U-H-F RECEIVER

The equipment was designed for two frequency ranges, 100 to 156 mc and 156 to 250 mc. The receivers were superheterodynes with two separate channels, one modifying the signals from the directional antenna and applying its output to the line transmitter, the other amplifying the nondirectional antenna signal and furnishing output to a headset for aural monitoring and a voltage for automatically controlling the volume of both channels. A common heterodyne oscillator served both purposes.

Provisions were made for handling either c-w or pulse signals; the i-f bandwidth could be set at 250 kc for c-w or at 3.5 mc for pulses by switching transformers in four of the six i-f stages. The circuits were designed to handle pulses having a repetition rate of from 625 to 4,000 per second and having a pulse width of from 1 to 15 μsec.

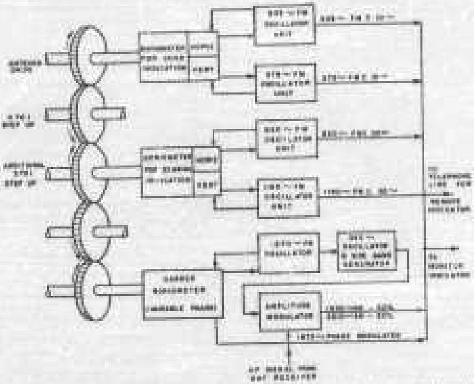


FIGURE 2. Block diagram of line transmitter showing several frequencies employed to transmit information to telephone lines.

14.5 LINE TRANSMITTER GONIOMETER UNIT

This unit obtained electrical information as to the exact and instantaneous position of the antenna and prepared these signals and the

output signal of the d-f channel of the receiver for transmission over the telephone line. Three goniometer assemblies were required, each geared through a differential to the rotating antenna. One goniometer rotated at the same speed as the antenna, providing X and Y components for tracing out the angular position of the antenna on the quadrant-indicating CR

A total of seven audio signals was used to transmit this information to the line receiver. Frequencies and amplitudes of these signals were proportioned to produce the least amount of distortion and crosstalk in the telephone lines. A block diagram of the line transmitter showing several frequencies employed to transmit information is given in Figure 2.

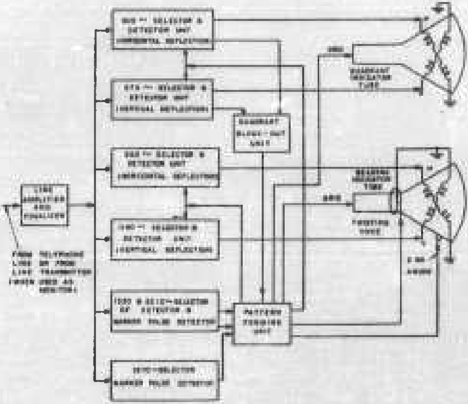


FIGURE 3. Block diagram of line receiver and CRO indicator tube.

tube, another rotated at four times the antenna speed and produced the components for tracing out the angular position of the antenna on the bearing-indicator CR tube, and the third goniometer rotated at 12 times the antenna speed for producing phase-modulated signals for electrical markers on the bearing-indicator cathode-ray tube.

14.4 LINE RECEIVER AND CATHODE-RAY INDICATOR UNITS

The line receiver (Figure 3) separated and prepared the signals received from the line-transmitting unit as to antenna location and d-f signal output for tracing the polar diagrams on cathode-ray tubes, one for indicat-

ing the directional lobe of the received signal for approximately locating the signal and another bearing-indicator cathode-ray tube having an expanded scale such that one complete revolution on the screen was equivalent to 90° of antenna rotation. On this tube a portion of the directional lobe was also traced out.

Because the lobe itself was not sharp enough to indicate the bearing accurately, circuits were provided for switching a deflection field at a rapid rate so that two intersecting patterns appeared on the face of the tube. The point of intersection of these patterns enabled the operator to determine azimuth accurately.

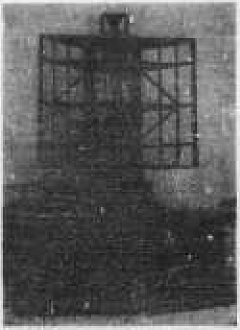


FIGURE 4. View of [illegible] direction finder with [illegible]

Electrical markers at 1° intervals with distinguishing marks at 5° and 15° intervals were provided. Transient traces were blocked out so that clear patterns were obtainable. A quadrant blockout control blocked out any two quadrants, a useful feature when examining two signals of the same frequency. The signals that were blocked out were shown electrically on the quadrant indicator tube by dotted traces. Only solid traces shown on the quadrant tube were reproduced on the bearing-indicator tube. A sample indicator pattern is given in Figure 5.

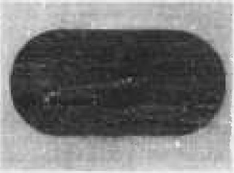


FIGURE 5. Sample pattern obtained when taking bearings.

Provisions were made for equalizing the telephone circuits. A pre-emphasis control was available for use where a Signal Corps line was connected between the d-f station and a telephone line, enabling the input to the Signal Corps line to be increased so that the signal arriving at the commercial facilities had the proper level.

14.7 APPARATUS LIMITATIONS

Effective service was accomplished on signals having strengths of 50 μv per meter or less. More modern techniques would enable this figure to be increased by a factor of five or more. The automatic volume control in the d-f channel of the u-h-f receiver obtained its voltage from the monitor channel so that the gain of the d-f channel was controlled in proportion to the input level of the monitor channel. Inasmuch as the monitor signal was not exactly constant as a funtion of antenna rotation, it was necessary to have a reasonably long time constant (approximately ½ second) in the a-v-c circuit for the d-f channel so that minor fluc-

tuations resulting from antenna rotation would not distort the d-f pattern and cause bearing error. Hence the d-f channel automatic volume control would, in general, respond to only relatively slow changes in signal level. Rapid changes caused a proportionate distortion in the d-f pattern which were indicated as instantaneous bearing errors on the cathode-ray screen. Such rapid variations caused the indicated bearing to vary about the true azimuth. Averaging the bearings of several traces visually enabled the operator to obtain the correct bearing.

Chapter 15

ELECTRICAL DIRECTION-FINDER EVALUATOR

Development of an electromechanical device which, from the bearings to a radio transmitter measured by any number of fixed radio direction finders, determines the most probable location of the transmitter and the boundary of the smallest region in which, to any preassigned probability, the transmitter can be presumed to be located.

15.1 INTRODUCTION

At the time of this project[a] there were, in use or available, a great number of radio direction finders capable of providing information which, if properly analyzed statistically on simultaneous bearings, could determine the location of a radio transmitter with much greater precision than had been obtained by methods of evaluation then existing.

This report describes a device which, without mathematical approximations and almost instantaneously, can apply the method of least squares to the bearings of any number of direction finders operating in a network. In conjunction with d-f networks organized to make optimum use of its properties, this electrical d-f evaluator was expected to place direction finding in an entirely new category of precision and dependability.

15.2 STATEMENT OF PROBLEM

A radio direction finder provides means for measuring the bearing to the source of a radio signal, and therefore two direction finders can provide sufficient information to determine the position of a radio transmitter, provided that the position of the transmitter is not on the line joining the two direction finders.

The bearings from the two direction finders will determine a fix (point where the bearing lines cross) with an accuracy dependent upon the precision of the two direction finders. In common with all physical measurements, the bearings as obtained from a direction finder deviate about the true value. And as with all physical measurements, if a number of values will be obtained and properly averaged, a resultant value will be obtained more dependable than any of the individual values.

The use of a number of direction finders instead of only two will provide information which, if properly averaged, will determine the location of a transmitter with greater precision than would the bearings from any two of them. In fact, the bearings from a large enough number of instruments can provide information for a fix of any desired accuracy. But the difficulty is in properly averaging the bearings. Unlike the measurements, for example, of the temperature at some location by a number of thermometers whose readings can be averaged by determining a simple mean value, the correct bearing of a transmitter from each direction finder of a network is in general a different value, and thus the mean value of the several bearings from direction finders located at different positions has no significance. If a method for correctly averaging their readings is used, the accuracy of a d-f fix is theoretically limited only by the number of direction finders. In Appendix A of the final report[1] the theory is fully expounded.

15.3 VISUAL D-F EVALUATION

The method usually employed in averaging the information obtained from a number of direction finders is to plot the bearings on a map of the region involved, and then, by visual observation, to estimate on the map the most probable location of the transmitter. This process makes use of various rules-of-thumb, geometrical constructions, and common-sense approximations in an attempt to obtain the coordinates of the most probable location of the transmitter. The more direction finders there are in a network, the less likely is the result of visual evaluation to approach the correct solution of the proper averaging process. The other

[a] Project 13-121, Contract No. OEMsr-1472, J. A. Maurer, Inc.

desired value: the boundary of the "search region," that is, of the smallest region in which to any preassigned probability the transmitter can be presumed to be located, cannot even be estimated by the visual evaluation method commonly employed. And yet this information may be very important in certain situations, such as, for instance, upon the reception of a distress signal, when the size and shape of the area most profitably to be searched by rescue craft should quickly be determined.

15.4 GROUP D-F SYSTEM OF EVALUATION

Another method to average the values from several direction finders has been attempted. This requires that a number of direction finders be located so close together that in effect they may be considered to have the same geographical location yet they must be far enough apart to prevent electrical coupling and to allow the errors in each instrument to be entirely uncorrelated. Thus if half of the direction finders are grouped at one location and half at another, the bearings within each group may be averaged by simply computing the mean value, and the resulting two bearings are used to obtain a fix on a map as if each were from a single direction finder, except that each mean bearing should be more precise than a bearing from a single direction finder. As experimentally tested, this group d-f system has been disappointing. Aside from the obvious limitation of having only two locations, it was found that when several direction finders were placed close enough to be treated as at one geographical point (not more than 2 miles apart) the deviations were not statistically random, and so in other words the errors were correlated, and the mean value of the bearings taken by a group was not much more dependable than the bearing from one direction finder alone.

15.5 USE OF THE SUMS OF THE SQUARES OF THE DEVIATIONS

The requirements for properly averaging the bearings from a number of separated direction finders may be represented geometrically in Figure 1 where the dotted lines represent the reported bearings from three direction finders as plotted on a map of the region, and the solid lines represent assumed bearings which meet in a common point T. The angles b_1, b_2, and b_3

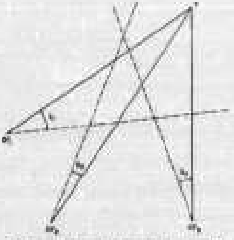


FIGURE 1. Location of transmitter by three d-f stations, dotted lines representing reported bearings, solid lines being assumed bearings which meet in common point T.

are the deviations between the reported bearings and the assumed bearings to the common point T. If the deviations of each direction finder are normally distributed (this is described in Appendix A of the final report[1]), then the most probable location of the transmitter is that position of T for which the sum of the squares of the deviations is a minimum.

A method which has been developed for evaluating d-f fixes analytically on a map comprises a series of computations of the sums of the squares of the deviation angles. In the neighborhood of the estimated location of the fix, a number of points in regular pattern are marked. By means of a transparent protractor, the deviation angle of each point from the reported bearing line of each direction finder is measured. These angles are then squared and added together for each of the points. The resulting values of the sums of the squares computed for each point give an indication of where the minimum sum would be located if an infinite number of points were used.

[illegible] BASIC PRINCIPLES OF THE ELECTRICAL D-F EVALUATOR

The electrical d-f evaluator does not use any approximations nor are any computations required during the actual evaluation process. Instead, it provides a mechanism whereby the common point T of Figure 1 can be moved to any position and simultaneously a reading proportional to the sum of the squares of the deviation angles is indicated on an electric meter. Thus by varying the position of T until the sum of the squares of the angles of deviation is a minimum the most probable location of the transmitter can be determined. The contour which bounds the smallest region in which to any preassigned probability the transmitter can be presumed to be located may also be determined from the sums of the squares of the deviations. Each curve along which this sum is a constant value (indicated by a constant meter reading) is a contour of constant probability density for the location of the transmitter. For any number of direction finders and any desired probability a value of this sum may be determined. Actually, the value of the sum has been computed for various probabilities and is provided with the evaluator in the form of a table.

In the development of the evaluator, tests were run on various d-f networks which verified the requirement that deviations of direction finders are approximately normally distributed, and thus the method of least squares is proper for these determinations.

FIGURE 2. Electrical direction-finder evaluator.

The above description assumes that the bearings as reported from each direction finder are equally dependable. In case it is known that the several direction finders have unequal precisions, the deviation angles (b_1, b_2, and b_3

of Figure 1) are weighted by dividing each by the standard deviation of the respective instrument. In the evaluator the weighting is performed in a circuit in which the squares of the deviation angles are measured, and therefore, the weighting control is a measure of the *variance*, which is the square of the standard deviation.

18.7 BASIC MECHANISMS OF THE ELECTRICAL D-F EVALUATOR

The electrical d-f evaluator, illustrated in Figure 2, is approximately the size and shape of the visual evaluating tables now in use by the Army Airways Communication System and the U.S. Coast Guard. It performs the operations of determining the minimum value of the sum of the squares of the weighted angles of deviation by means of a number of protractors located at points representing the positions of the direction finders on a gnomonic chart of the region involved. Each protractor electrically measures the square of the angle between the reported d-f bearings and the great-circle line to the common point T of Figure 1. In the evaluator, this point can be manually moved to any position on the map, and is called the scanning point. Each protractor is a form of potentiometer carrying 60-cycle alternating current and is constructed basically of a resistance strip attached to the bearing disk which can be oriented to the azimuth reported by the direction finder, and a wiper attached to a telescoping pointer arm which leads to the scanning point. From each protractor a separate pointer arm leads to the same scanning point. The resistance strip and wiper of the protractor are so designed that a voltage is obtained proportional to the square of the angle measured by the relative position of the pointer arm from the reference line on the bearing disk. External to the protractor is a selector switch which permits the 60-cycle current to each protractor resistance strip to be so regulated that the voltage from each can be weighted according to the variance of the direction finder representing the protractor. The voltage from each protractor is applied to the primary of one of a bank of "summation transformers." The secondaries of these summation transformers are in series, and the series output is applied to the grid of a vacuum-tube amplifier whose amplification is variable in five steps. The output of this amplifier actuates the "summation meter," and this is the meter whose reading is proportional to the sum of the squares of the weighted deviation angles.

18.8 PANTOGRAPHS

Because each protractor is a fairly large component (about 4 in. in diameter) and because direction finders are occasionally located rather close together, it would not be practical to place all the protractors side by side on a chart. In the electrical d-f evaluator this difficulty is resolved by providing a number of decks, permitting the different protractors to be located at different levels, but each is directly below the point on the gnomonic chart representing the position of its corresponding direction finder. The scanning point appears as the reference point with a marking pencil at the end of a movable arm just above the map on top of the evaluator structure, but at each deck of the evaluator there is a duplicate scanning point attached by a horizontal pantograph and vertical shaft assembly to the scanning point so that it always remains directly below it. It is to the duplicate scanning points that the pointer arms from the various protractors are pivoted.

18.9 GNOMONIC CHART DISTORTION CORRECTION

The chart or map used with the evaluator must be a gnomonic projection because only with such a projection are all great circles represented by straight lines. This projection, however, has one property which presents difficulties in measuring the angles of deviation at various parts of the map. This property is called nonconformality and because of it angles on the surface of the earth are not preserved in the flat projection.

To overcome this difficulty, a corrector assembly is employed in each protractor by which the angle between the wiper and the resistance strip is altered by a cam to compensate exactly the gnomonic distortion.

13.10 D-F BEARING INPUT

At the right end of the evaluator is a series of boxes called bearing-input boxes, one for each direction finder of the network. Each contains an internally illuminated translucent drum with an engraved scale reading 0-360° which can be rotated by means of a 36/1 ratio bearing knob to the bearing reported by the corresponding d-f station. A flexible shaft runs from the bearing knob of the bearing-input box to the bearing disk of its associated protractor through a 36/1 ratio worm so that drum and protractor rotate together.

Each bearing-input box also has a 10-point switch by which the current to the resistance strip in the protractor can be varied to provide proper weighting for deviation angles according to the known dependability of the particular direction finder. The weighting is dependent upon the statistical history of each direction finder.

Means are provided for visual evaluation in case a breakdown occurs of the electromechanical system.

13.11 METHOD OF OBTAINING THE DESIRED DATA

When all the reported bearings have been entered into the bearing-input boxes and the variance switches are at their proper positions, the operator moves a vertical pencil on the end of a pantograph arm above the map. With one hand controlling the sensitivity of the summation amplifier, the pencil is moved until a minimum is notad on a summation meter. A mark is made on the map at this point. Then the pencil is moved perpendicularly to the first straight line and a new motion described parallel to the first line and a mark made when a new minimum is found. Now on a line joining these two points a third minimum will be found. It will be very close to the most probable location of the transmitter. The pencil may be caused to describe short motions about this point to find an absolute minimum and this will locate the most probable location of the transmitter.

Means are provided for rejecting "wild" bearings. In the contractor's final report[1] are given a procedure for describing the boundary of search regions of any given probability, and statistical data resulting from field tests on east and west coasts; also the report gives consideration to further developments of the electrical-evaluator circuits, directions for making the cams, the use of servo mechanisms to eliminate the manual manipulation of the protractors, and to means of making the computations required automatic.

13.12 ACKNOWLEDGMENTS

In the design and development of the electrical d-f evaluator, certain individuals and groups in the Armed Services of Great Britain and the United States rendered considerable assistance.

The theoretical and practical requirements for an improvement in d-f evaluation were originally presented to Division 13 in great detail by Captain Stuart Martin, Office of Chief Signal Officer, U.S. Army. The results of his considerable research on statistical methods in d-f evaluation were generously provided by Commander D. H. Menzel, Op. 20G, U.S. Navy.

The long-term statistical studies of d-f station errors and group d-f station experiments carried on in Great Britain by Crampton and Redgment were made available, together with valuable interpretive information, by Admiralty Signals Establishment.

The U.S. Army Airways Communications System and the Air Sea Rescue Section of the U.S. Coast Guard cooperated continuously through their headquarters, their training centers, and their several d-f evaluation offices in providing equipment, operational data, special records, and experimental information at all stages of the development.

Kenneth A. Norton and Ross Bateman, attached to the Office of Chief Signal Officer, U.S. Army, were, through the later stages of development, in such close cooperation with the designers that certain features of the evaluator are directly attributable to them.

The final report was prepared jointly by personnel of Division 13, NDRC, by personnel of the Applied Mathematics Group, and by J. A. Maurer, Inc.

PART III

RADIO AND WEATHER

Chapter 16

A STUDY OF SFERICS

The work on this project[a] was divided into two parts. The first was a survey of existing literature on the subject of atmospherics and their relation to weather information; the second consisted in the operation of two radio stations in New Mexico in cooperation with the Signal Corps to gather visual, electrical, meteorological, and photographic data on local thunderstorms. While the contractor submitted completion reports[1,2] covering both phases of the project, the summary following is condensed only from the one[2] covering the experimental operations.

16.1 INTRODUCTION

THE PURPOSE of this project was to gather as much data as possible on thunderstorms and the types of sferics (atmospherics) they produced with the object of answering the following questions.

1. Can thunderstorms be located accurately?
2. Given a distribution of thunderstorms, can the weather situation be analyzed?
3. Are there characteristics of sferic signals which can be associated with storms of definite type or energy which will supplement or clarify the information obtained from geographical distribution of storms?
4. In any given region do thunderstorms occur with such frequency that the sferic direction-finding technique can be profitable?

The project consisted of two parts, a survey of the pertinent literature available and an exploratory experimental program. Only the experimental program is described herein.

16.2 EQUIPMENT UTILIZED

Two observing stations were set up, one at the University of New Mexico in Albuquerque and one on top of the Sandia Mountains. The Signal Corps provided a mobile-unit-equipped sferic-waveform and d-f apparatus which was located at various distances from 80 to 1,500 km from the University station. The observational data on lightning flashes were synchronized with the sferic records in the mobile unit by means of radio signals. The Signal Corps also provided waveform and d-f apparatus for use at the University and a B-17 plane with equipment similar to that in the mobile unit. The plane was not continuously available during the time of the project.

Each station was equipped with an electrical potential gradient change recorder consisting of an exposed insulated electrode connected to a quartz string electrometer and to ground through a high resistance.[3] The time constant of the system was chosen so that gradient changes due to lightning strokes occurring within a few hundredths or tenths of seconds produced large electrometer deflections but slow gradient changes of seconds' duration produced no deflections. The gradient changes (electrometer deflections) were recorded on a 16-mm film moving at constant speed past a slit 0.002 in. wide. The instruments were sufficiently sensitive to record gradient changes due to lightning strokes within a radius of 25 miles and fast enough to resolve gradient changes due to repeated elements of lightning flashes.

Each station also was provided with a tape recorder on which the time, type, and azimuth of lightning flashes and the time of the thunder were recorded. Frequent time signals and lightning stroke signals were keyed on the gradient change recorders and simultaneously transmitted by radio to the mobile unit to synchronize the several records. In addition, each station was equipped with an alidade to measure storm and lightning flash azimuth and cloud base and top elevation angles.

Time lapse photographs of cloud development were taken from each station.

[a] Project 13-115, Contract OEMsr-1485, University of New Mexico.

Mobile Unit

The sferic d-f equipment in the mobile unit consisted of AN/GRD-1 apparatus[4] made up of two square loops mounted at right angles for detecting perpendicular components of the incoming signals. The separate amplifiers were properly phased and the component signals impressed on the horizontal and vertical plates of a cathode-ray tube. The sets were tuned to a frequency of approximately 10 kc.

The sferic waveform equipment consisted of a vertical 36-ft antenna, an aperiodic antenna circuit, an amplifier with nearly constant amplification up to about 200 kc, a cathode-ray tube, and a triggering circuit. The latter started the sweep after the sferic was received with a delay of about 5 μsec. The amplified sferic was impressed on the vertical plates so that the cathode-ray tube trace represented the field variations of the sferic signal with time. The time base or sweep used varied between 1,500 and 2,000 μsec. The sweep was calibrated by impressing 10-kc or 20-kc sinusoidal signals of various amplitudes on the apparatus. A block diagram of the d-f and waveform equipment is given in Figure 1.

Both the d-f and waveform scopes were photographed simultaneously by a 35-mm camera. A signal lamp for synchronizing signals was mounted between the scopes. The film moved continuously at a rate of approximately 2 in. per second.

OBSERVED WAVEFORMS AND STORM DISTANCE

The waveforms observed can be divided into three principal types.

1. A series of prominent, easily distinguished features (e.g., maxima, sharp breaks), usually with amplitudes decreasing in a fairly regular fashion forming a repeated pattern. The usual sweep with 1,800-μsec time base showed from two to five such features. The interval between the features characteristically increased from 100 to 200 μsec at the beginning of the trace to 400 to 600 μsec at the end of the trace.

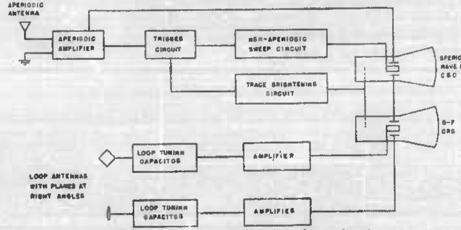


FIGURE 1. Block diagram of d-f and waveform equipment.

2. A series of prominent features with less regular intervals and greater amplitude variation than in the first type. The waveform frequently suggested an interference pattern formed by two or more superimposed pulses or oscillations.

3. Very complicated waveforms with varying amplitude and with intervals between maxima from 10 to 100 μsec.

Waveforms of Type 1 were analyzed according to the suggestions of Laby[7] and Schonland[8] on the assumption that the pulses or oscillations

were due to multiple reflections between earth and ionosphere. According to this hypothesis the time of transit of an electromagnetic disturbance from a lightning stroke to the observing station is

$$t_n = \frac{1}{c}\,(4n^2h^2 + d^2)^{1/2}$$

where c is the velocity of propagation of the disturbance;
h is the height of the ionosphere;
d is the great circle distance between source and observer;
n is the number of reflections at the ionosphere experienced by the pulse.

The time between the arrival of a pulse which has been reflected at the ionosphere n times and one which has been reflected $n - 1$ times is

$$\Delta t_n = \frac{1}{c}\left\{(4n^2h^2 + d^2)^{1/2} - \left[4(n-1)^2h^2 + d^2\right]^{1/2}\right\}.$$

In the analysis of the sferic waveforms, the procedure was to choose distinguishable repeated parts of the pattern (maxima, minima, sharp breaks, etc.), measure the time intervals between them, and calculate h and d by the above formula.

for by multiple reflections from an ionosphere 90 km in height suggesting a storm to the east where the path of the sferics would be in the dark. The largest concentration of directions lay between 80° and 90° azimuth with a maximum at 85°. The calculated distance of the sources was 1,375 ± 100 km. The storms producing the sferics were thus located within 120 km of the center of Arkansas. Weather data of the date showed that a number of thunderstorms occurred along a cold front extending from Arkansas to Pennsylvania. At the time the records were made a storm was in progress at Little Rock, Arkansas. Thus the location of the storm at this site without previous knowledge of its existence on the part of those analyzing the records offers convincing evidence of the validity of the multiple reflection hypothesis.

Lightning Flashes and Storm Character

A study of the visual and electrical potential gradient change records of lightning strokes in storms near Albuquerque during August and September, 1945, yielded some interesting preliminary results. In this group of storms, the frontal storms were more intense, they had a

Table 1. Results of waveform analysis.

Date	Time of obs of correlated flashes, MST	Storm distance from mobile unit d in km	Calculated ionosphere height h in km	Path of sferic	No. of corr. waveforms consistent with h and d	No of waveforms corr. by time only	No of waveforms corr by time and direction
Aug 30	1938-1946	400	90	Dark	3	4	4
Aug 31	1437-1500	590	75	Light	21	23	21
	1513-1515	580	78	Light	2	2	2
	1543-1550	600	80	Light	5	7	7
	1626-1630	600	82	Light	10	11	11
Sept 6	1853-1915	840	85	Dark	4	5	4
		Total		...	45	52	49

The results obtained by this means of analyzing the waveforms are given in Table 1.

The record of the storm of November 2, 1945 disclosed many simple patterns which, upon analysis, indicated that they could be accounted

greater stroke frequency, a relatively larger number of cloud-ground strokes, and a larger number of repeated elements per stroke than intra-air-mass storms. If these observations are supported by further studies over entire thunderstorm seasons and in different climatic regions, there is a possibility of determining

storm types as well as storm distance from sferics waveform records.

16.8 GENERAL CONCLUSIONS

1. Sferic signals from lightning flashes experience multiple reflection between the ionosphere and earth. The repeated pattern waveform produced by sferic pulses which travel paths of different length due to different numbers of ionospheric reflections may be used to calculate the height of the ionosphere and the distance of the flash from the observing station.

2. The use of waveform equipment to determine lightning flash distance in conjunction with equipment to measure direction and sense of the sferic signal makes possible location of thunderstorms from a single station. A thorough test of this technique should be made.

3. Preliminary results on a small group of thunderstorms in one climatic region indicate that frontal and nonfrontal storms differ in lightning flash frequency, relative number of cloud-ground and cloud-cloud flashes, number of repeated elements in cloud-ground flashes, and the duration of cloud-ground flashes.

4. The great advantage of determining storm type or intensity from sferics records indicates that the preliminary results should be checked and extended by observations on storms in several climatic regions.

PART IV

ANTENNA RESEARCH

Chapter 17

ANTENNA PATTERNS FOR AIRCRAFT

Studies and experimental investigations in connection with antenna patterns for aircraft and tanks as a function of location of the antenna, frequencies employed, etc., also development of the "model" technique for studying aircraft antenna impedances and patterns. This contract was administered by Division 13 until April 1, 1944, when it was transferred to Division 15.

17.1 INTRODUCTION

PROJECT C-11[a] was initiated by NDRC at the request of Aircraft Radio Laboratory, Wright Field, to achieve the following principal aims.

1. To investigate methods for measuring antenna patterns on aircraft at various frequencies.

2. To measure the patterns of various antennas on various types of aircraft at various frequencies.

3. To obtain general statements on the effects of aircraft structure, antenna location, frequency, and other factors on the radiation patterns.

4. To investigate the patterns of various special antennas and antenna arrays.

5. To investigate methods for improving patterns of aircraft antennas for specific applications.

6. To investigate the construction of models to determine the accuracy of construction required.

17.2 RESULTS ACCOMPLISHED

Although measurements of aircraft patterns using models had been made for several years prior to the start of this project, the measurements were limited to simple types of antennas and to an upper frequency of about 500 mc. Under the project, techniques and equipment were developed to extend the model methods to a greater variety of structures and to cover greater frequency ranges. After the equipment and techniques had been developed to the point where routine measurements could be made, at frequencies as high as 10,000 mc, patterns of various antennas were investigated to determine the general factors which influence the patterns. It was found possible to predict the general features of patterns of simple types of aircraft antennas.

Modeling techniques were applied to a variety of special problems and it is believed that these applications are new. Methods for measuring propeller modulation and for measuring ellipticity of polarization of aircraft antennas were developed. Modeling techniques were applied in the investigation of a tank antenna problem. The possibility of using models for measuring radar echoes from aircraft was considered and development of methods started. Methods using models for measuring the impedances of aircraft antennas were investigated.

The research program outlined above was requested by Wright Field in order to develop the model technique for use as a tool in the design of aircraft antennas to meet definite specifications. Models were used in the investigation in preference to full-scale aircraft since they furnish more information with less labor, time and cost. The fact that the actual airplane is not always available for antenna tests also was an important factor.

The information and techniques developed on this project were used in the design and development of aircraft antennas for a wide variety of applications.

[a] Project C-11, Contract No. NDCrc-100, Ohio State University.

17.3 PATTERNS OF ANTENNAS ON AIRCRAFT AT VARIOUS FREQUENCIES

It is not easy to predict from theoretical considerations alone the approximate patterns to be expected from a proposed antenna installation on an airplane. The relative importance of reflection and diffraction effects and the nature of the current distributions on the surfaces of the aircraft are difficult to estimate. If sufficient antenna patterns measured under a wide range of conditions are available, it becomes possible to make a better estimate of an antenna pattern. To provide such patterns, a group of patterns has been obtained over a wide frequency range for simple antennas mounted on various types of aircraft.

Only the patterns for the principal planes have been measured. It has been found that principal plane patterns are almost as useful as complete three-dimensional patterns, and much easier to obtain. The orientations of the coordinate planes with respect to the aircraft are shown in Figure 1.

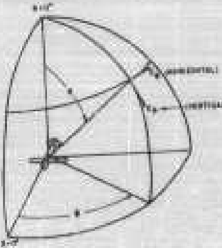


FIGURE 1. [illegible]

The following is a list of the patterns included with the contractor's final report dated August 24, 1943.[1]

B-17F

A 4-ft whip antenna on the lower frequencies and a $\lambda/4$ stub at 15, 25, 35, 50, 75, 100, 150, and 200 mc, the antennas being located (1) directly ahead of the bomb bays, projecting vertically downward, (2) directly behind the bomb bays, projecting vertically downward, (3) 4 ft ahead of the leading edge of the horizontal stabilizer, projecting vertically downward from the belly of the ship, and (4) centered on wings on top of fuselage, projecting vertically upward.

A-20-A

A $\lambda/4$ stub antenna on top of the fuselage, immediately above the trailing edge of the wing at 50, 100, and 200 mc.

P-38

A 4-ft whip antenna projecting forward from the nose at 50, 100, 150, and 200 mc.

P-47

A 4-ft whip antenna just behind the pilot's cockpit at 50, 100, and 200 mc.

B-25

Two types of antennas, a $\lambda/4$ stub and a $\lambda/2$ coaxial-type dipole at 100 and 200 mc. The antennas were mounted in two locations, on top of the fuselage, first just above the leading edge of the wing, and then above the trailing edge.

17.4 TYPICAL ANTENNA PATTERNS

In making the measurements only half of the pattern was measured in those cases where symmetry could be assumed. The symmetry was checked in several of the patterns and found to be adequate.

In Figures 2 and 3 the row of patterns on the left is for the plane $\theta = 90°$, the center row for the plane defined by $\phi = 0°$ and $180°$, and the right-hand row for the plane $\phi = 90°$ and

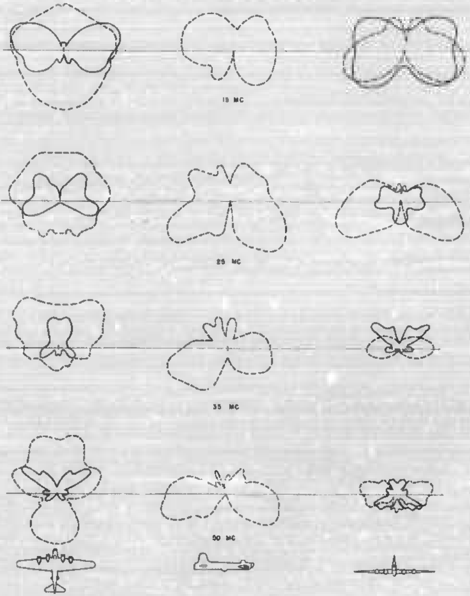


FIGURE 2. Antenna patterns for 4-ft stub on belly of B-17F directly in front of bomb bays. Dotted lines indicate vertical polarization, E_θ; full lines indicate horizontal polarization, E_ϕ.

270°. In Figures 4 and 5 the patterns for angles 10° below the horizon ($\theta = 100°$) have been plotted also.

The principal plane patterns in any horizontal row in Figures 2 and 3 are plotted on the basis of a constant power input and there-

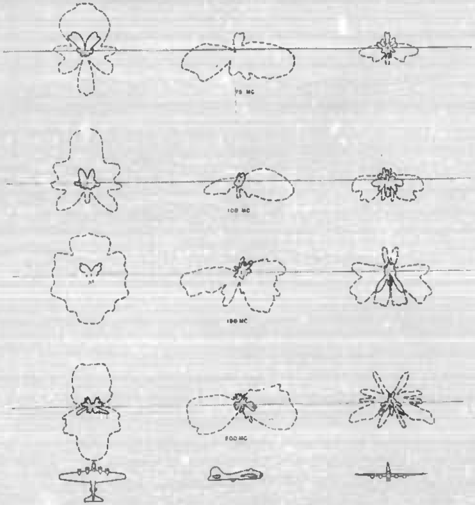


FIGURE 3. Antenna patterns of λ/4 stub on belly of B-17F directly in front of bomb bays. Dotted lines indicate vertical polarization, E_θ; full lines indicate horizontal polarization, E_ϕ.

fore may be directly compared. It is not permissible to make direct comparisions of relative signal strengths between patterns in different rows.

The pattern of any simple antenna mounted on an airplane may be estimated with the aid of these sample patterns. The sample patterns which approximate the conditions of the antenna whose pattern is to be estimated are compared to determine the amount of diffraction and reflection to be expected. If the current distribution on the antenna is expected to differ considerably from that obtaining on the stubs used in these measurements, due allowance for its effects on the pattern must be made. It will be found, however, that the

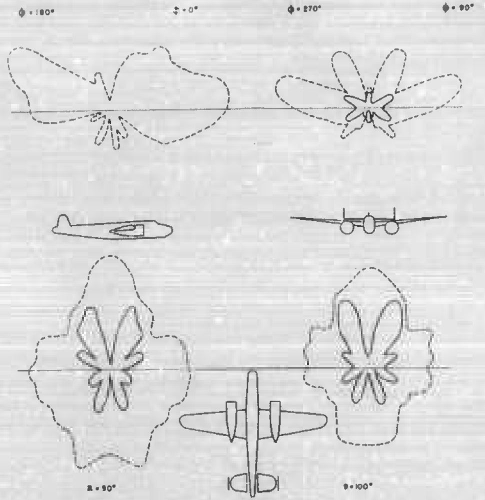


FIGURE 4. Patterns of $\lambda/4$ stub at 100 mc on top fuselage above leading edge of wing of B-25. Dotted lines indicate vertical polarization, E_θ; full lines indicate horizontal polarization, E_ϕ.

sample patterns will be approximately correct for linear antennas of lengths from a small fraction of a wavelength up to roughly $\frac{5}{8}\lambda$.

As an additional aid in estimating antenna patterns, a number of patterns were measured on a $\lambda/4$ stub mounted on a prolate spheroid, which approximates a fuselage. It is apparent from the patterns in Figures 6 and 7 that their shapes are determined more by the nature of the current distribution on the spheroid than by the current distribution on the antenna.

17.8 METHOD OF MEASUREMENT EMPLOYED

A fairly adequate description of the principal methods employed in measuring antenna

patterns with models is given in the final report dated August 31, 1942.[1] A few minor changes were made as a result of experience. The vibrator method described briefly below has certain advantages over other methods especially in certain applications. The fact that no connecting wires to the model are required is of particular advantage in some measurements. The phasing adjustment offers possibilities for investigating the ellipticity of polarization of radiation from an antenna. On the other hand, the vibrator method has certain disadvantages.

1. The amount of modulation obtainable with a commercial vibrator is very low at frequencies about 2,000 mc due to unavoidable stray reactances and losses in the vibrator.

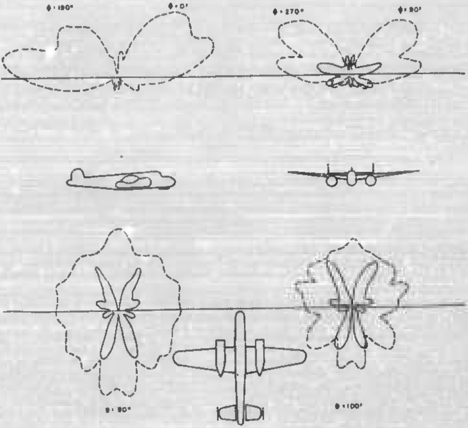


FIGURE 5. Patterns of λ/4 stub at 200 mc directly above leading edge of wing of B-25. Dotted lines indicate vertical polarization, E_θ; full lines indicate horizontal polarization, E_ϕ.

2. The need for phasing the system for each reading increases the time required to measure a pattern compared to other methods. It is possible, probably, to eliminate this phasing adjustment.

3. The signal levels obtained are low, and the system is rather sensitive to changes in components.

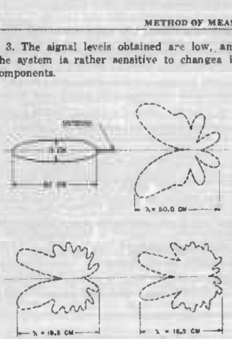


FIGURE 6. Patterns obtained from λ/4 stub projecting from one end of prolate spheroid 15x60 cm in dimensions, vertical polarization.

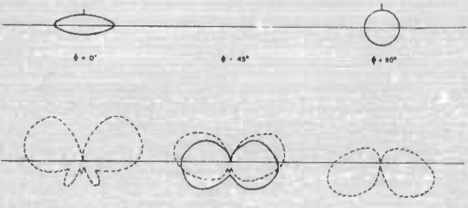


FIGURE 7. Patterns of λ/4 stub on side of prolate spheroid, 15x60 cm in dimension, parallel to minor axis; vertical polarization.

17.3.1 The Vibrator Method

An unmodulated transmitter produces a relatively uniform field in the region occupied by the model exciting the model antenna. The current which flows in the model antenna is modulated by connecting a periodically varying impedance (tuned vibrator) to the terminals of the antenna. As a consequence of the variations in antenna current, a modulated wave is re-radiated. Some of the re-radiated energy re-enters the transmitting antenna system where it is picked up by a receiver sensitive to modulation only. Since there are two signals entering the receiver, the audio output of the receiver depends upon their relative phase. The phase may be varied by adjusting the separation between the model and the transmitting antenna. Variations in the adjustment for proper phasing (maximum audio output) yield information on phase variations in the field re-radiated from the model.

17.3.2 The New Method

The method employed for the majority of the pattern measurements uses a bolometer (Littlefuse) detector as a receiver in the model to detect modulated signals from a horn radiator. Small wires are used to connect the output of the receiver to the observing position. Provided suitable precautions are taken, the distortion of the antenna pattern due to the presence of these wires in the field can be kept small. For antennas of low efficiency, the out-

put of a bolometer is rather low so that a silicon crystal detector is usually substituted.

The output of most detectors is essentially proportional to the square of the input voltage. Since antenna patterns are usually plotted on a voltage basis (to accommodate the large variations in signals found in most patterns) it is necessary to take the square root of the voltage output of the receiver. An amplifier which does this automatically has been constructed. It is essentially a logarithmic 50-kc amplifier whose components have been adjusted to give the desired square root characteristic.

The model supporting structure described on page 34 of the final report dated August 31, 1942,[1] is now used exclusively. Selsyn indicators give a remote indication of the rotational position of the horizontal member. The horn radiator is on rollers to allow complete freedom of rotation about its longitudinal axis.

17.4 PATTERNS OF BALANCED ANTENNAS

Patterns of antennas requiring a balanced feed cannot be measured as simply as those using a coaxial-feed system. Particular care must be taken to assure a balance in the currents on the feed line otherwise stray currents appear on the outer shield, distorting the measured pattern.

Since the measuring equipment was originally designed for use with coaxial lines, the first method used on balanced antennas employed a

Figure 8. Coaxial skirt balancing unit.

$\lambda/4$ skirt or balancing section on the end of a coaxial line to obtain the phase reversal required for a balanced antenna. (See Figure 8.) This method has several disadvantages, the most important being the necessity for changing the length of the skirt with each frequency change. Also, the length of the skirt is quite critical if the antenna impedance is high. It is often difficult to find space in a model for the matching section.

A modification of this method is shown in Figure 9. A sliding polystyrene plug inserted in

Figure 9. Tunable coaxial skirt balancing unit.

the skirt unit allows some adjustment of the tuning of the skirt. The tuning range is rather restricted, however, and there is no good criterion for proper tuning.

The next method tried used a balanced system throughout. Shielded-pair transmission lines and balanced detectors were constructed, as shown in Figure 10. Two coaxial tuners

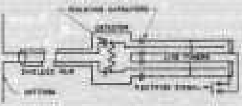


Figure 10. Balanced measuring system with separate adjustments for balance.

were used at the detector to allow adjustment of balance since the detectors were not quite symmetrical mechanically. This method was found to be satisfactory for a wider range of antenna impedances than the previous methods. There was still a lack of a criterion for proper tuning, however.

A system which achieved greater mechanical

and electrical symmetry is shown in Figure 11. A twin-line tuner and dual detectors were used. The two bolometers were connected in series for the audio output. This equipment was relatively satisfactory.

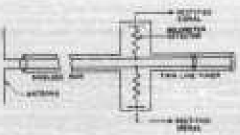


Figure 11. Balanced system using dual detectors.

A system which uses a resonator to couple an unbalanced detector to a balanced transmission line is shown in Figure 12. This avoids the difficulties encountered in constructing balanced detectors.

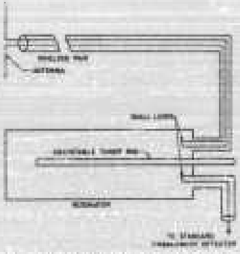


Figure 12. Use of resonator for coupling balanced line to coaxial line.

17.7 PROPELLER MODULATION

Preliminary tests were made to determine the feasibility of using models to study propeller modulation. For a given direction of propagation of the signal, it is possible to observe the variation in signal when the propeller is oriented in various directions. From the maximum and minimum signals observed it is possible to determine the percentage of modulation due to the propeller.

17.8 MEASUREMENTS ON ELLIPTICALLY POLARIZED FIELDS

Radiation from even simple stub antennas mounted on aircraft is elliptically polarized at the higher frequencies. It is to be expected, therefore, that measurements of the ellipticity of the radiation would yield information of value in interpretations of patterns.

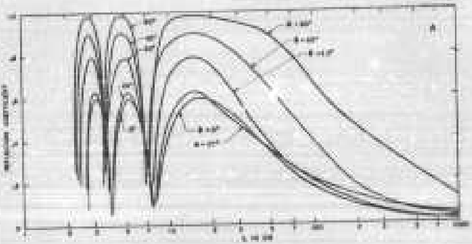
The major and minor axes of the ellipse of polarizations at any given point in a field can be readily measured by rotating a linearly polarized antenna to determine the maximum and minimum signals. If the field is linearly polarized the minimum signal will be zero. If the field is circularly polarized there will be neither a maximum nor a minimum. To determine the direction of rotation of the electric vector around the ellipse special measurements are required. The phasing adjustment used in the vibrator method for measuring antenna patterns makes its determination possible.

Measurements have been made of the ellipticity of the field radiated from a simple vertical stub antenna located to the rear of the cockpit of a P-40 at 150 mc. The data obtained are tabulated in Table I of Appendix I (report dated August 31, 1942).[1] Table II[1] was obtained from measurements of the field radiated from a $\lambda/4$ stub antenna located on the side of a prolate spheroid parallel to a minor axis of the spheroid. There is a considerable amount of elliptically polarized radiation in directions not in the planes of symmetry. The direction of rotation of the polarization was not measured in the pattern for the P-40.

17.9 THE SIMULATION OF DIELECTRICS IN MODELS

An accurate simulation of a dielectric in a model is obtained by using a material whose dielectric constant is the same and whose cor-

ductivity has been increased by the factor by which the dimensions have been reduced. Since suitable materials were not readily available, an investigation was conducted to determine type aircraft, is in some cases important. Also, in certain special cases it is necessary to model plastics such as Plexiglas. The enclosures used on some antennas, such as loops and high-fre-



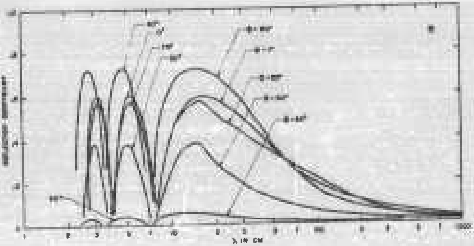


FIGURE 13. Reflection from 2-cm layer of wood. A, horizontal polarization; B, vertical polarization. ($\epsilon = 4$; $\sigma = 50 \times 10^{-11}$ emu.)

methods for constructing approximate models for pattern measurements. The modeling of plywood, such as is used for constructing cargo-quency radar antennas, sometimes affect the pattern of the antenna.

The precise calculation of the effect of a

curved dielectric surface, such as a plywood fuselage, on the propagation of waves radiated from an antenna is difficult and involves too much labor to be practical. Much useful information is obtainable, however, from the simpler calculation involving plane waves and plane surfaces.[1]

The reflection coefficient for a 2-cm layer of plywood was calculated for a number of angles of incidence and for both vertical and horizontal polarization. The values for dielectric constant ϵ and the conductivity σ were obtained by averaging published values from a number of sources. At the time of the calculations the [illegible] of Roberts and Von Hippel[2] was not available. The results of the calculatio[illegible] are shown in Figure 13. An examination of these figures shows that there is negligible reflection for wavelengths longer than about 10 meters. As the wavelength is decreased below 10 meters the reflection increases to a maximum in the region around 15 cm. Beyond 15 cm the reflection coefficient exhibits alternate maxima and minima, the plywood acting as a pure dielectric reflector.

For antennas which operate at wavelengths longer than 10 meters the plywood may be expected to have but small influence on the antenna pattern. Consequently it is not necessary to model the plywood at all. It will be necessary to model any conducting materials in the field of the antenna, such as the motors and gas tanks.

For the region from 15 cm to shorter wavelengths, a reasonably accurate model may be obtained using plywood of proper thickness in the model, since the conductivity becomes unimportant.

For the region between 15 cm and 10 meters, the situation is not so favorable. Reflections from the surfaces of the aircraft may have considerable influence on an antenna pattern. The model should be constructed of materials having the correct constants if accurate results are desired. An approximation to the pattern may be obtained by using a plywood model, and the results will usually be good enough to indicate the general performance of the antenna system. The errors in the pattern will depend on how much the waves reflected from the plywood surfaces contribute to the antenna pattern.

17.10 TANK ANTENNA PATTERNS

The following investigation was undertaken to determine a method for measuring the patterns of certain high-frequency antennas mounted on a medium tank. It was considered necessary to include the effect on the patterns of the finitely conducting ground in the neighborhood of the tank.

17.11 THE MODEL TECHNIQUE

The characteristics of the ground on which a tank is located may influence the pattern of an antenna on the tank in two ways. The most important effect at high frequencies is the change in the pattern due to reflections from the surface of the earth. Of lesser importance, generally, is the effect of the ground on the current distribution on the antenna and on the tank.

An electromagnetic wave incident on a surface of finite conductivity and dielectric constant is ordinarily reflected with a change in magnitude and phase and possibly a change in polarization. The wave received at any point in space from an antenna near the earth's surface will be the vector sum of a direct wave plus a wave reflected from the surface. Because of the change in phase on reflection and because of the difference in path traversed, the phase difference between the direct and reflected waves at a point in space will depend on the relation of the point to the antenna.

An accurate simulation of the constants of the ground could be obtained for model measurements by using a model ground constructed of a material whose dielectric constant equals that of the earth and whose conductivity is increased by the factor by which the dimensions in the model are reduced. Suitable materials of these characteristics were not readily available, although there was a possibility of obtaining them by loading rubber with a large amount of carbon. Mechanical difficulties in the model equipment made it desirable to obtain the patterns by other means if possible.

The antennas of principal interest operated at frequencies sufficiently high so that it could reasonably be assumed that the presence of

the ground had only a negligible effect in determining the current distribution on the antenna and on the tank. If, therefore, the model tank is suspended in free space, it can be as-

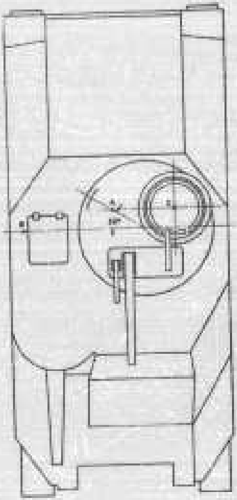


FIGURE 14. Top view of tank showing locations of antennas.

sumed that the current distribution is unchanged. It thus becomes possible to measure the pattern of the current distribution in free space. From theoretical considerations, an estimate of the pattern including the effect of the ground can then be made.

Free-space pattern measurements were made on a stub-type antenna mounted on a medium tank in two positions (positions marked A and C in Figure 14). Photographs of three-dimensional models of the patterns appear in Figures 15 and 16. The patterns in Figure 16 show the influence of the position of the turret on the pattern, since the turret is not located symmetrically on the tank. The frequency used in these measurements was such as to make the height of the tank about 5λ.

FIGURE 15. Three-dimensional pattern of antenna mounted on small turret of tank.

The free-space patterns can be modified to obtain an approximation to the true pattern including the effect of ground reflections. An examination of the pattern in Figure 15 shows that there is very little energy radiated at angles more than about 20° below the horizon, owing to the large surfaces of the tank body. Only those waves included in the region from

the horizon to 20° below the horizon can therefore be expected to influence the pattern after reflection from the ground. Since the angle of reflection is equal to the angle of incidence, only the pattern in the region up to 20° above the horizon will be distorted by the ground reflections. For angles greater than this, the measured patterns are probably about correct.

For angles up to 20° above the horizon, the measured pattern must be modified to include the ground reflections. By multiplying the average magnitude of the signal in the region 0° to 20° in the measured free-space pattern by the pattern for the corresponding region for the dipole, an approximation to the true pattern is obtained. The calculated pattern for an infinitesimal dipole located 5λ above an average earth is shown in Figure 17. Figure 18 shows a pattern which has been modified in this way.

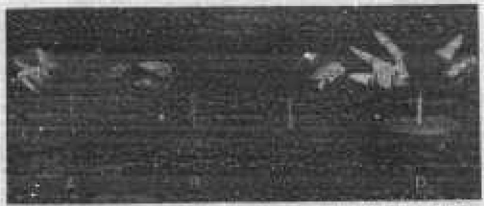


FIGURE 16. Patterns for antenna mounted on large turret. A, guns pointing to rear; B, guns pointing to right; C, guns pointing left; D, guns pointing forward.

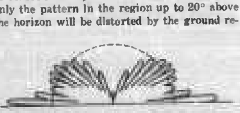


FIGURE 17. Pattern of vertical dipole approximately 5λ above average ground.

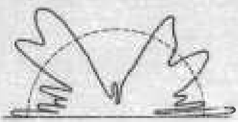


FIGURE 18. Pattern of vertical stub antenna on small turret.

17.12 PATTERNS FOR A CURVED ANTENNA ON A MEDIUM TANK

An examination of the patterns in Figures 15 and 16 shows that there is a severe cone of silence along the axis of the antenna. As this cone of silence might be deleterious in the case of tank-to-plane communications, a curved antenna was investigated to see if the cone of silence could be eliminated. Measured free-space patterns indicated that a more uniform distribution of signal was obtained with this antenna. While there is some energy in the horizontally polarized component, it was not important enough to warrant measurement. The presence of the ground will have about the same effect on the pattern of this antenna as on the pattern for the stub discussed above. The exact shape of the antenna to produce the uniform distribution of signal is not too important, provided only that an appreciable component of the axis of the antenna is horizontal.

17.18 IMPEDANCE MEASUREMENTS BY MEANS OF MODELS

In designing an antenna for a specific application there are generally two electrical factors which have to be considered, the pattern and the impedance characteristics. For many applications it is possible to design the antenna to produce the desired pattern and to accept whatever impedance characteristics result. There are many applications, particularly broad-band antennas, where it is not permissible to choose the one characteristic independently. To be able to correlate the pattern with the impedance characteristics it was felt desirable to attempt measurement of antenna impedances by means of models. After a review of the principal methods described in the literature for measuring impedances at ultra-high frequencies, the standing-wave method and a modification of Chipman's method offered most promise of being adaptable to the problem.

Equipment was constructed for making standing-wave measurements in the frequency range 600 to 3,000 mc. The first crude equipment revealed the necessity for very precise mechanical construction. Deviations of the center conductor from the axis of the outer conductor as small as 0.001 in. caused very noticeable distortions in the standing wave pattern.

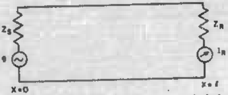


FIGURE 19. Basic circuit of Chipman's method of measuring impedance.

Most of the development of methods has centered on Chipman's method, for which the basic circuit is shown in Figure 19. A sketch of the equipment used is shown in Figure 20. The antenna whose impedance is to be measured is connected to one end of a coaxial transmission line, and excited by a remotely located transmitting antenna. Varying the length of the line to obtain a resonance curve supplies data from which the unknown impedance can be determined. An indication proportional to the current in the short-circuiting plunger of the transmission line is obtained by means of a small coupling loop and transmission line to a detector. The plunger is driven by a micrometer drive. A silicon detector and galvanometer are used for the detector.

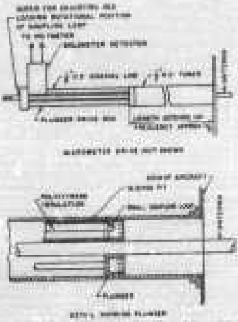


FIGURE 20. Measuring equipment used in [illegible]

The calibration of the equipment was carried out as follows. Owing to the dissymmetry introduced by the plunger, there is some uncertainty as to the location of the origin from which the length of the line is measured. A calibration was obtained by short-circuiting the antenna end of the line with a disk, and determining the resonance position for the frequency used by feeding energy into the detector line. The resistance introduced by the detector-coupling loop (it was assumed that

there is no reactance introduced since the line is tuned) was obtained from the resonance curve in the usual way with no antenna connected to the line. Energy was fed into the open end of the line by means of a probe introduced near the open end.

Measurements have been made of the impedance of a vertical rod antenna mounted on a large plane conductor at 750 mc. Measurements were made of the impedance as a function of the length of the antenna. The procedure was simply to determine the position of the plunger for resonance and the breadth of the resonance curve at the half-power points.

The results of the measurements are shown in Figures 21 and 22. A curve calculated from

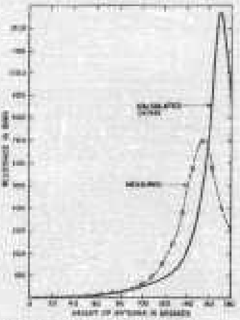


Figure 21. Resistance of cylindrical rod antenna. [illegible]

the formula given by King and Blake[3] has also been plotted for comparison. The large deviation from the calculated values has been found by other experimenters, and so is not considered serious.

17.14 THE RECIPROCITY BETWEEN TRANSMITTING AND RECEIVING ANTENNAS

Since some radio engineers have doubts concerning the validity of the reciprocity theorem for antenna patterns,[4] it seemed desirable to make a rough test of it. The basis for this doubt lies in the feeling that known differences in current distribution on an antenna when transmitting and when receiving should lead

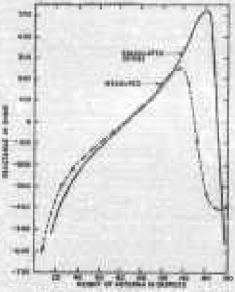


Figure 22. Reactance of rod antenna of Figure 21.

to differences in pattern. The opinion has been expressed that the nature of the impedance at the terminals of a receiving antenna might affect the pattern.

A rough test of the reciprocity theorem was made by measuring the patterns of a number of antennas when receiving and when transmitting. In no case was any attempt made to keep the impedances of the generator or the current indicator negligible, as required by the reciprocity theorem as usually stated. The agreement between the receiving and trans-

mitting patterns was satisfactory in all cases. The antennas tested included $\lambda/2$ linear symmetrical antennas, and a flat antenna mounted on a disk.

The method which has been described for measuring antenna impedences based on Chipman's work, measures the impedance of the antenna when receiving. It is, therefore, important to show that this impedance is the same as when transmitting. The proof is essentially that given by Franz.[4] Consider any antenna with an impedance Z connected to its terminals and under the influence of an arbitrary incoming wave. The wave sets up a current I_r in the impedance Z. Suppose now that a generator is inserted in series with Z to reduce the current flowing in it to zero. The generator sets up a current I_t which is equal and opposite to I_r. By the superposition theorem, I_r is unaffected by the presence of I_t, and vice versa. Hence,

$$I_r + I_t = 0 \tag{1}$$

but

$$I_t = \frac{E_r}{Z + Z_a} \tag{2}$$

where Z_a is defined as the impedance of the antenna when transmitting. It follows immediately that the receiving antenna acts like a generator of open-circuit voltage E_r and internal impedance Z_a. Hence the impedance of a receiving antenna is identically the same as that when transmitting. It is apparent that Thevenin's theorem applies to a receiving antenna.

Note that the equality of the currents expressed in the equations applies only to the currents in the impedance Z. It does not imply that the currents are equal over the entire antenna, since in general they will not be equal except at the terminals.

Chapter 18

AIRBORNE ANTENNA DESIGN AT U-H-F AND V-H-F

Study of antenna design taking into account directional properties, power-handling capacities, propagation at u-h-f and v-h-f frequencies, horizontal and vertical polarization, broad-band antennas, surface antennas, effect of the structure on drag, precipitation static, etc. The summary which follows is condensed from the contractor's final report,[1] the chief reduction consisting in the elimination of a great number of field patterns, drawings, and bibliographical references.

18.1 INTRODUCTION

THE OBJECT OF this project[a] was to "study optimum radiation patterns for use on or in aircraft in the v-h-f and u-h-f ranges from a communication point of view, taking into account the variety of attitudes in which an airplane may be when communication may be necessary, to study designs of antennas required to realize such optimum radiation patterns, and to include consideration of aerodynamic aspects to the end of attaining antennas presenting minimum drag." The work resulted in a report[1] containing as complete information as could be obtained within the specified time.

In view of the extremely general nature of these specifications and the short time allowed, it was not considered feasible to undertake special experimental or theoretical work. Therefore, this summary and the final report[1] from which it is condensed are necessarily based upon work done for or by the various agencies of the Armed Services. The bulk of the work was performed by:

1. The Antenna Section, Research Division, Aircraft Radio Laboratory, Wright Field, Dayton, Ohio.
2. The Radio Test Department, U. S. Naval Air Station, Patuxent River, Maryland.
3. The Robinson Laboratory, Ohio State University Research Foundation.
4. The Radio Research Laboratory [RRL], Harvard University.
5. The Radiation Laboratory [RL], Massachusetts Institute of Technology.
6. The Bell Telephone Laboratories [RTL], New York City and Deal, New Jersey.
7. RCA Laboratories [RCAL], Rocky Point, Long Island.

[a] Project 13-105, Contract No. OEMsr-1396, Radio Corporation of America.

18.2 GENERAL CONSIDERATIONS IN AIRBORNE ANTENNA DESIGN

Since it is a rare antenna installation in which the transmitter or receiver can be located directly at the antenna terminals, the problem is usually complicated by the presence of a transmission line. In practice, with low-loss transmission lines of essentially real characteristic impedance, the power transfer problem is solved by so designing the transmitter and the antenna that their respective input impedances are resistive and equal in value to the characteristic impedance of the line. Under these conditions the line is said to be "flat" or "matched," the energy delivered to the line passing down the line in the form of a traveling wave, which, on reaching the antenna, is entirely absorbed and radiated into space.

When the antenna is not matched to the transmission line, the incident voltage wave is reflected at the antenna terminals with a change in magnitude and a shift in phase determined by the input impedance of the antenna relative to the characteristic impedance of the line. The effect of such reflection is to set up a system of standing waves on the line, the characteristics of which are described, for engineering purposes, in terms of two equivalent quantities: the magnitude of the reflection coefficient $|K|$, and the standing wave ratio SWR. These are defined in terms of the terminating impedance and the voltage distribution on the line by the following expressions:

$$|K| = \left|\frac{Z_0 - Z_t}{Z_0 + Z_t}\right|, \tag{1}$$

where Z_t is the terminating (or antenna im-

pedance) and Z_0 is the characteristic impedance of the line,

and $$SWR = E_{max}/E_{min}, \qquad (2)$$

where E_{max} and E_{min} are the relative magnitudes of the maximum and minimum voltages in the standing wave system.

The two quantities $|K|$ and SWR are related by the expressions:

$$|K| = \frac{SWR - 1}{SWR + 1} \quad \text{and} \quad SWR = \frac{1 + |K|}{1 - |K|}. \qquad (3)$$

10.3.1 Resonant Lines

It is evident from the general transmission line equation (for lines of real Z_0 and negligible attenuation)

$$Z_{in} = Z_0 \frac{jZ_0 \tan \theta + Z_t}{Z_0 + jZ_t \tan \theta} \qquad (4)$$

that while the input impedance Z_{in} of a flat or matched line is independent of θ, the electrical length of the line, and always equal to the characteristic impedance of the line, the input impedance of a mismatched, or resonant, line is quite definitely a function of line length. If, at a given frequency, an antenna is mismatched to an extent described by a given SWR, the input impedance of the line is determined by the line length through equation (4), the resistive and reactive components R and X of the impedance assuming any values satisfying the equation

$$\frac{(R^2 + X^2 + Z_0^2)}{RZ_0} = \frac{(SWR^2 + 1)}{SWR} \qquad (5)$$

which is that of a circle (of constant SWR) in the complex impedance (R-X) plane.

The effect of this dependence of input impedance upon line length is that while a given transmitter may be adjusted to work directly into a given mismatched antenna over a range of frequencies, the same transmitter may refuse to work into a low-loss line terminated in the same antenna. It is necessary to minimize the SWR by matching the antenna, if it is desired that a transmitter deliver energy to the line over an appreciable range in frequencies without special adjustment, the range and rate of variation of input impedance being greater the greater the SWR and the longer the line.

10.3.2 Effect of SWR on Line Voltage

For a given power delivered to the antenna terminals, the maximum line voltage is greater the greater the SWR. The greater the voltage the greater the possibility of line failure due to arc-over, particularly at connectors and other discontinuities in the line. Dielectric losses, other things being equal, are proportional to the square of the line voltage; the presence of high SWR on a solid-dielectric line is often indicated by local heating effects in the vicinity of voltage maxima, particularly near the transmitter end of the line. Furthermore if the line voltage is high and a voltage maximum happens to fall at a line discontinuity the effect of that discontinuity will be greater the higher the SWR; many an otherwise satisfactory antenna system has been impaired by the unfortunate location of a cable connector with respect to the standing wave system. Figure 1

Figure 1. Effect of standing wave ratio (SWR) on power transfer into lossless line.

shows the effect of SWR on power transfer at the antenna terminals, for the mythical case of a lossless line.

10.3.3 Transmission Line Losses

Losses in the polythene-filled flexible coaxial cable ordinarily used in aircraft antenna installations are of two general types: resistive or "skin-effect" losses in the cable conductors, and dielectric losses in the polythene. These losses

introduce attenuation according to the following expressions:

$$\alpha_c = 13.6\frac{d}{\lambda}\left(\frac{1+\frac{b}{a}}{b}\right)\frac{\sqrt{\epsilon}}{\log_e\frac{b}{a}} \tag{6}$$

and

$$\alpha_D = \frac{27.3\sqrt{\epsilon \tan\delta}}{\lambda} \tag{7}$$

where α_c is the attenuation in db per cm due to conductor losses.
α_D is the attenuation in db per cm due to dielectric losses.
d is the skindepth in cm.
λ is the wavelength in air in cm.
b is the radius of the dielectric in cm.
a is the radius of the inner conductor in cm.
ϵ is the dielectric constant of the cable dielectric.
$\tan\delta$ is the loss factor of the cable dielectric.

The sum of these two attenuations is the total attenuation (α_T) of the line. Since skin depth varies inversely as $\sqrt{f}$, conductor losses increase as $\sqrt{f}$; dielectric losses increase directly as the frequency. Therefore conductor losses are more important at low frequencies, dielectric losses becoming more serious at high frequencies; this effect is shown in the following table applying to RG-14/U coax, a medium-power cable in common use in aircraft radio.

Attenuation in RG-14/U cable

Frequency in mc	α_c in db/100 ft	α_D in db/100 ft	α_T in db/100 ft
100	1.04	0.21	1.35
3,000	5.66	6.23	11.9

The effect of line attenuation on overall efficiency may be made evident by the fact that as little as 25 feet of RG-14/U cable has sufficient attenuation at 3,000 mc to reduce the maximum efficiency to less than 50 per cent—even if the antenna and the transmitter are perfectly matched. For this reason the use of appreciable lengths of solid dielectric cable is avoided at frequencies in the upper u-h-f range, wave guides being used instead if efficiency is required.

18.8.4 Transmitting Antenna Characteristics

Since the weight and power capacity of airborne transmitters are severely limited, the necessity for a low *SWR* on the transmission line imposes rather rigid restrictions on the characteristics of the transmitting antenna. For the antenna to be efficiently matched to the characteristic impedance of the line over a wide frequency band, its impedance must be characterized by a resistance of the order of the line impedance, and by low and not too rapidly varying reactance. These conditions are most easily met in practice by antennas worked against the skin of the plane as ground, and operated in the vicinity of $\lambda/4$ resonance.[b] Nonresonant antennas, much less than $\lambda/4$ in length, do not make efficient transmitting antennas.

18.8.5 Electrically Short Transmitting Antennas

Vertical stub antennas worked against ground and less than $\lambda/4$ long have low radiation resistance and large capacitative reactance, the latter rapidly varying with frequency. If it were desired to use an antenna only 0.04 λ long for transmitting purposes, the stub, for a length/diameter ratio of 50:1, would have a resistance of about 1 ohm and a reactance of −1,000 ohms. While such an impedance could be matched to a 50-ohm line by means of a two-element transmission-line matching section at very high frequencies (where there is no particular point in using such a small antenna), at lower frequencies it could be matched in a practical way only by a matching section of lumped impedances. An L section would perform the double function of tuning out the 1,000 ohms capacitative reactance by means of a loading coil and of stepping up the 1-ohm resistance to look like 50 ohms. While such a matching section matches the antenna to the line, at the spot frequency in question, the frequency band over which the *SWR* is less than some reasonable maximum, say 2:1, is extremely small. Further-

[b] Throughout this report the symbol λ is used for wavelength; thus a $\lambda/4$ antenna indicates an antenna one-quarter wavelength long.

more, since the loading coil must necessarily have resistance (a Q of 250 has been assumed) only a fraction of the power entering the matching section will actually reach the antenna. In this example the transmitting system has a power efficiency that is at most 20 per cent, and even that small figure neglects transmission line losses and losses due to the ohmic resistance of the antenna and its ground system.

18.2.5 Characteristics of Good Transmitting Antennas

Efficient transmitting antennas are realized on aircraft in the form of $\lambda/4$ antennas worked against the metal skin of the ship, or antennas of $\lambda/2$ dipole-type suspended in space from the structural members of the plane. Even surface antennas, i.e., antennas mounted inside the skin of the plane and radiating through the apertures of slots, horns, and cavities, have critical dimensions of the order of $\lambda/2$ or more. Such antennas have high input resistance, of the order of the characteristic impedance of the feed line, and small reactance which is a relatively slowly varying function of frequency.

18.2.7 Antenna Impedance Measurements

Since the actual impedance characteristics of a practical v-h-f or u-h-f aircraft antenna rarely have more than a slight resemblance to theoretical impedance data, it is almost always necessary to determine these characteristics by actual measurement, if optimum performance is desired. This is particularly true with aircraft antennas for the lower v-h-f range and for any antenna located on surfaces of curvature comparable to the operating wavelength or near reflecting and resonating structures. In such cases the antenna impedance should be measured under conditions as nearly identical as possible with those under which the antenna is to be used in practice. The most satisfactory procedure, as far as results are concerned, is to conduct these measurements on the full-scale ship—in flight, if necessary—with the antenna complete in all details of its [illegible]al mounting and feeding system. Where this is impractical, a poor second-best procedure is to measure the antenna impedance by means of models, a $1/n$-scale model of the antenna being installed in proper location on a $1/n$-scale model of the plane and its impedance measured at n times the actual full-scale frequencies. This method is capable of good results only if great care is taken in scaling all details of the antenna and its mounting and feed system.

At higher frequencies, or in general for any aircraft antenna mounted on or in an airplane surface which constitutes a good approximation to a flat ground plane for at least λ in all directions, satisfactory results may be obtained by means of impedance measurements made with the antenna worked against a ground plane in an ordinary laboratory setup. But in all cases an effort should be made to ensure that the antenna is studied under conditions which closely approximate actual flight conditions.

18.2.8 Antenna Impedance Matching

While satisfactory antennas for some purposes can be realized without knowledge of the antenna impedance, by trial-and-error adjustment of tuning stubs and simple matching sections, modern methods of impedance matching presume a knowledge of the impedance characteristics of the antenna. These methods go far beyond the simple $\lambda/4$ transformers and shunt tuners described in texts and other published literature.

18.2.9 The Receiving Antenna Problem

The receiving antenna problem is different from that of the transmitting antenna problem in that the former is not so much concerned with power transfer but with the attainment of a high signal-to-noise ratio. This end can be approached in a twofold manner, by increasing the received signal strength and by reducing noise.

Noise may be picked up by the antenna along with the signal or may be generated in the receiver itself. In the upper h-f and lower v-h-f ranges antenna noise may be large compared to

that developed in the receiving circuit, but as the frequency increases antenna noise decreases, until at frequencies greater than 70 or 80 mc it is negligible compared to set noises. As far as the receiver proper is concerned, the attainment of high signal-to-noise ratio at ultra-high frequencies is largely a matter of reducing tube and circuit noise in a closer and closer approach to the limits set by thermal agitation.

There are many other sources of noise in aircraft radio reception since not only the antenna but the skin of the ship itself are parts of the receiving system. If poor electrical contacts exist anywhere in this system, the vibration associated with normal flight is likely to result in relative motion of the adjacent conductors at such contacts, which motion will appear in the receiver as noise. Hence the necessity for "bonding." Shielding is useful in reducing static of local origin. Antenna design features tending to minimize precipitation static are discussed elsewhere in this report.

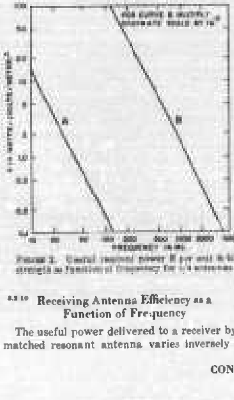


FIGURE 2. Useful received power [illegible] strength as function of frequency for [illegible] antennas.

8.2.10 Receiving Antenna Efficiency as a Function of Frequency

The useful power delivered to a receiver by a matched resonant antenna varies inversely as f^2, a fact which may severely limit the range of u-h-f communication. This relation is shown in Figure 2. While the received signal may be increased manyfold by means of arrays, horns, or reflectors, increased gain implies increased directivity, and extreme directivity is not usually a desirable feature in aircraft communication antennas. Furthermore, since the gain of a directive system is roughly proportional to its aperture area in square wavelengths, it is evident that, except at very high frequencies, a practical limit to the gain of an aircraft antenna is quickly reached.

18.2.11 Impedance Matching of Receiving Antennas

The effect of antenna mismatch is much less serious in receiving than in transmitting systems. Receivers may be designed to have input impedances equal to the characteristic impedance of the feed line over very wide frequency bands, and in such cases, even though the antenna may be very badly mismatched, there will be no standing waves on the line. The effect of mismatch is to reduce the signal reaching the receiver terminal, the loss in received signal voltage being a slowly increasing function of the degree of mismatch until the mismatch becomes quite large. Thus an antenna system which would be quite impossible for efficient transmission may well be very satisfactory for reception. For this reason the design standards for receiving antennas are usually much lower than those for transmitting, a simple stub or whip much shorter than a $\lambda/4$ often making a satisfactory antenna if the field strength is sufficiently high.

18.2.12 Effect of Line Losses on Line Input Impedance

The effect of line attenuation is to reduce the magnitude of the variation in input impedance of a mismatched line, with a resulting reduction in the apparent reflection coefficient or *SWR* looking into the line from the receiver or transmitter terminals. Thus the longer the line, and the greater its attenuation per unit length,

the flatter its input characteristic for a given degree of mismatch at the antenna. The effect of these losses is to make the loading of a transmitter or the tuning of a receiver a less critical function of frequency.

18.8.13 Reception of Very Weak Signals

In the u-h-f range, where the received signal is often not much greater than the receiver noise level, it may sometimes be found that a mismatch at the receiver will result in increased sensitivity. In such cases it is desirable that the receiving antenna be matched to the line, for otherwise standing waves can exist on the line, with multiple reflections resulting in multiple signals if line losses are low.

18.8.14 Radiation Characteristics of Airborne Antennas

In aircraft antenna design, particularly in the v-h-f range, it is necessary for the antenna to radiate or pick up energy in the desired directions in space. Unless certain fairly rigid conditions are met, there is usually not much resemblance between the actual field pattern of a given antenna on aircraft and that of the same antenna in free space or worked against a flat infinite ground plane. Except for special cases, which are most commonly met in practice only at u-h-f frequencies, it is necessary to demonstrate, by actual measurement, that the field pattern of a given installation is satisfactory for the application at hand.

A classic example of the absolute necessity for field pattern measurements is shown in Figure 3 where the diagrams represent the measured horizontal plane and vertical fore-and-aft-plane patterns of a 100-mc V-antenna installed on a 4FU. The antenna consisted of two wires strung outward from opposite sides of the fuselage, aft near the tail, to the tips of the horizontal stabilizer. The antenna was used, with rather disastrous results, at the start of the war in connection with an application which requires a pattern having a maximum or lobe in a generally forward and downward direction. This antenna actually had nulls in the important directions.

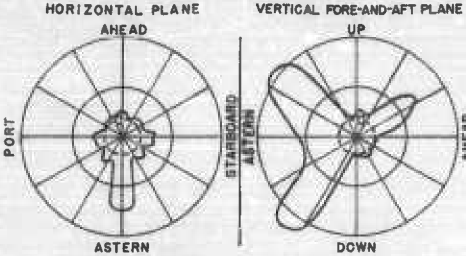


FIGURE 3. Two wires strung from sides of fuselage to tips of horizontal stabilizer, fed out of phase, horizontal polarization, 100 mc.

18.3 V-H-F AND U-H-F PROPAGATION

Plane-to-plane and plane-to-ground communication in the v-h-f and u-h-f bands is dependent almost entirely upon space-wave propagation, except for anomalies occurring under certain atmospheric conditions in which a sky-wave effect is introduced by "reflection" at the discontinuity between air masses of different synoptic properties.

18.3.1 The Space Wave

The space wave is not a single wave but rather the resultant of a direct or line-of-sight wave and a wave reflected by the ground. The direct wave is not direct in a strict geometrical sense, owing to refraction in the atmosphere and to diffraction around the bulge of the earth and around other obstacles. Under ordinary conditions the direct-wave field intensity is subject to little more than the inverse distance law attenuation of free space.

The ground-reflected wave is subject to all the laws of optical reflection and is ordinarily subject to greater attenuation than the direct wave, since the former must necessarily travel a longer path. For this reason, and because the magnitude of the reflection coefficient is less than unity except at grazing incidence, the amplitude of the ground-reflected wave at the receiver is less than that of the direct wave. Since the space wave is the resultant of two waves of different amplitudes and different phases, it is evident that it is a complicated function of elevation, distance, frequency, and polarization.

18.3.2 Ground-Reflection Coefficients

The nature of the ground-reflected wave is determined by the ground-reflection coefficient, which is different for vertical and horizontal polarizations. For horizontal polarization (the electric vector normal to the plane of incidence) the magnitude of the reflection coefficient drops steadily from unity to a smaller final value as the angle of incidence varies from grazing to normal incidence, while the phase of the reflection coefficient remains at substantially 180° for all angles from grazing incidence to normal incidence. For vertical polarization (the electric vector in the plane of incidence) the magnitude of the reflection coefficient drops rapidly from unity at grazing incidence to a small value at a small angle to the horizon, rising gradually with increasing angle to approximately equal to the value for horizontal polarization at normal incidence. Meanwhile the phase shift for vertical polarization decreases rapidly from 180° at grazing incidence to 90° at the angle of minimum reflection, finally becoming 0° for angles between the angle of minimum reflection and normal incidence.

Therefore the received signal in aircraft communication will vary with distance, elevation, and frequency.

18.3.3 Effect of Distance

At distances small compared to the antenna heights the resultant space-wave amplitude oscillates about its normal free-space value as transmission distance increases, since both the phase difference between the two component waves due to their different path lengths and the phase difference due to the fact that the ground-reflection coefficient is a function of angle of incidence depend upon distance. With increasing distance these oscillations develop larger amplitudes, since the amplitudes of the two component waves approach equality as their path lengths become more nearly equal and since the magnitude of the ground-reflection coefficient approaches unity at grazing incidence. But while the amplitude of the oscillations increase, their frequency decreases with distance, since the increment in distance for a given phase difference becomes greater with increasing distance.

At distances large compared with the antenna heights, corresponding to ground reflection at grazing incidence, but still above the line-of-sight horizon, the received signal is no longer oscillatory, but obeys almost exactly the inverse-square law.

At still larger distances the component waves approach phase opposition and the resultant

field drops off more rapidly than the normal free-space field as the distance increases beyond the optical horizon. The phenomena described above are very similar for both horizontal and vertical polarization, except in the near region in which the distance is of the same order of magnitude as the elevation of the antennas; in this region the oscillatory resultant field is more complex for vertical polarization owing to the greater sensitivity of the vertical reflection coefficient to angle of incidence, when that angle is large.

18.3.4 Effect of Elevation

The effect of elevation is quite similar to that of distance in the near region in which the height is not negligible compared to the transmission distance; the same oscillations in received signal occur in the field strength versus height curve as in the field strength versus distance curve, and for the same reasons.

At greater distances, beyond the oscillatory region, the field strength is almost proportional to the product of the antenna heights. Below the optical horizon the field strength is very sensitive to antenna height. For low heights field strength is at first independent of height, then increases with altitude at an accelerated rate until it is directly proportional to height. At very great elevations (but still below the optical horizon) field strength increases more rapidly with height than as the product of the antenna heights.

18.3.5 Effect of Frequency

As far as the oscillatory region is concerned the effect of increasing frequency is to increase the number of oscillations per unit distance and to extend the distance over which oscillations occur. This effect is particularly pronounced with horizontal polarization over ground, the extent of the oscillatory region increasing from approximately 5 to approximately 100 miles as the frequency is increased from 30 to 600 mc (for the case of communication between a ground station and a plane at 40,000 ft altitude).

At greater distances, but still above the horizon, an increase in frequency results in an increase in field strength for horizontal polarization; for vertical polarization there is no significant change. Below the optical horizon (into which region the space wave extends by virtue of refraction and diffraction) the field strength is less the higher the frequency, particularly for vertical polarization.

For all of these reasons dependable v-h-f and u-h-f communication is restricted to stations above each other's optical horizon. Although this situation precludes the possibility of extremely long-range communication, it is not nearly so severe a restriction as might be expected, since plane-to-plane and plane-to-ground communication can extend over quite respectable distances.

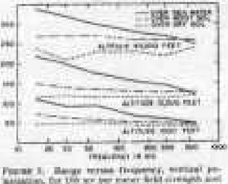
Figures 4 and 5 summarize the effects of frequency, polarization, distance, elevation, and nature of the ground upon high-frequency propagation. While these curves are based on calculations for ideal short doublet antennas and take no account of the field pattern of an actual aircraft antenna, they are valuable in that they give a qualitative picture of how the controlling factors affect aircraft communications.

FIGURE 4. [illegible]

18.3.6 Polarization and Propagation

Figures 4 and 5, and the preceding discussion indicate little difference between vertical and horizontal polarization, as far as propa-

gation over land is concerned. In the near region vertical polarization is preferable, since the field strength is greater and the magnitude of the oscillations less pronounced. Vertical polarization has better transmission over sea water and moist soil at the lower frequencies.



[illegible] 10 w radiated power, one antenna [illegible] the other at altitudes indicated, short doublet antennas. (Data from Bell Telephone Laboratories.)

[illegible] Anomalous Effects at High Frequencies

Tropospheric Reflection

Since the velocity of propagation of u-h-f radio waves in air depends upon the dielectric constant of the atmosphere, which in turn is a function of pressure, temperature, and humidity, it is natural that there be a correlation between anomalous propagation and the passage of meteorological "fronts," and with the existence in the upper atmosphere of any abnormal distribution of temperature and humidity. Under such conditions, as the radio ray passes into the region of different electrical properties it will be refracted, perhaps sufficiently to return to earth, giving rise to what are known as tropospheric reflections, although they are not reflections in the strict optical sense, since the discontinuity is not sharply defined.

Trapping or Waveguide Effect

Such reflections may make communication possible over much greater distances than are ordinarily attained. If the discontinuity layer is sufficiently pronounced the radiation may be effectively trapped at the top of the inversion layer, the region between this layer and the surface of the earth acting somewhat like a waveguide having large attenuation. While this waveguide effect may be helpful in making extremely long distance communication possible, it may also be a liability at smaller ranges, depending upon the height of the inversion in the stratified atmosphere, due to interference between the ordinary space wave and this pseudo sky wave.

Fading at U-H-F

This interference results in fading, which may vary with time as well as with altitude and distance, due to the relative motion of the two air masses resulting in shifting of the position of the discontinuity layer.

Fading and trapping are generally more pronounced the higher the frequency, partly because directive antennas are usually used for both transmission and reception at the higher frequencies; and since the sharper the radio beam the greater the fraction of energy reflected by the discontinuities, and the more apparent their effects.

At present there is no conclusive evidence that either polarization is less affected by atmospheric disturbances.

Static at V-H-F and U-H-F

The higher frequencies are much less affected by static of natural origin than are low. Due to the absence of a sky wave under normal conditions, u-h-f communication is less susceptible to static from distant sources. Since the field strength of static is approximately proportional to wavelength, static of local origin is less effective the higher the frequency.

[illegible] Man-Made Interference

Noise generated in rotating machinery and other sources of interference seems to be pre-

dominately vertically polarized; consequently such interference is generally worse for vertical aircraft antennas than for horizontal.

18.5.9 Multipath Interference

Since the wavelengths corresponding to very-high and ultra-high frequencies are small compared to the dimensions of buildings, hills, etc., there may be multiple ground-reflected rays resulting in even more complex interference effects in space-wave communication than those discussed above.

Furthermore, the structural members of the aircraft upon which the antenna is mounted are large enough relative to small wavelengths to cast shadows and cause reflection and diffraction effects which may interfere greatly with transmission and reception in certain directions. These effects are generally more pronounced the higher the frequency.

18.5.10 Propeller Modulation

Propeller modulation, at a frequency equal to the product of the number of propeller blades by the number of revolutions per second, will affect both transmission and reception. Although such modulation can approach 100 per cent in extreme cases, it can be minimized by removing the antenna from the immediate vicinity of the motors.

18.6 FIELD PATTERNS OF ANTENNAS ON AIRCRAFT

Since the radiating system formed by an aircraft antenna and the skin of the plane upon which it is mounted is generally quite complex, the field patterns of such a system are usually quite different from those of a similar antenna in free space or mounted on a flat infinite conducting plane. Except under certain special conditions, usually met in practice only in the case of u-h-f antennas mounted on or in flat, unobstructed airplane surfaces of dimensions large in terms of wavelength, experience shows that the field pattern of a given antenna will be modified to a greater or less extent by the plane upon which it is used. To be certain that the field pattern of a given aircraft antenna installation satisfies the requirements of the problem, it is usually necessary to determine the field pattern experimentally.

18.6.1 Flight Measurements of Field Patterns of Aircraft Antennas

The most direct method for determining the radiational characteristics of a given antenna on aircraft is to install the antenna on the plane upon which it will be used, connect it to a transmitter covering the frequency range in question, and fly the plane in a definite course around a field-strength meter located on the ground. While direct, this method has disadvantages: it is a difficult procedure, requiring that the plane be flown on a prescribed course maintaining constant speed, distance, and elevation; there are perhaps less than a dozen pilots in the country with sufficient skill and practice to make accurate pattern measurements possible. Furthermore flight measurements are expensive and time-consuming, often—in times of plane and personnel shortages—an outright impossibility.

The most serious objection to flight measurements is that at best they yield information about only a very small part of the total field pattern of an aircraft antenna. Because of the oscillatory nature of the space-wave field upon which aircraft communication depends it is difficult, if not impossible, to obtain meaningful field-strength measurements when the test airplane is at distances comparable with its elevation. The oscillatory region extends from 5 to 100 miles from the ground station, depending upon elevation, frequency, polarization, and ground conditions; accordingly the angular spread of the space pattern that can be measured in flight with any pretense at accuracy is extremely limited—0 to 10° below the horizon being an optimistic range. While this range of elevation angles could be extended by banking the plane, the pattern pilot usually has enough to do without having to maintain his plane at a constant angle of tilt. The use of a second

plane to replace the ground station would merely multiply the difficulties already present and would introduce the further complication of the directivity of the receiving antenna system on that second plane.

For these reasons flight measurements are usually restricted to the determination of field pattern in the horizontal plane of the ship.

18.4.2 Pattern Measurements by Means of Models

Model measurements, based upon the principles of similitude and reciprocity, form the most satisfactory method for determining the radiation patterns of aircraft antenna systems. In this method an accurate $1/n$-scale model of the plane is mounted on a nonmetallic tower in such a way that the model has two degrees of rotational freedom and is located in the uniform field of a pyramidal horn radiating energy of frequency n times that used on the full-scale plane. The model plane is remote from the horn (in terms of wavelength), the directivity of which is such that there is no danger of interference effect due to reflection from the ground or from nearby obstacles. A $1/n$-scale model of the antenna is mounted upon the metallic surface of the plane in its proper position and is connected, through appropriate feed and matching systems, to a thermocouple, bolometer, crystal, or other detector, located inside the model, the d-c output of the detector being fed from the model to a remotely located microammeter or other indicating or recording device. The reading of the d-c instrument bears some simple relationship to the r-f signal received by the antenna. By properly orienting the model plane, whose position with respect to the horn is usually remotely controlled, it is possible to determine the relative field pattern of the antenna in any or all directions in space.

The validity of model pattern measurements depends largely upon the accuracy with which the model and the antenna are constructed and scaled. It depends also upon the measuring equipment, particularly in regard to frequency stability and constancy of output of the oscillator and upon uniformity of the field pattern of the horn over the entire region in space in which the antenna can be situated as a result of the motion of the model. Aside from the care required in scaling the length of the antenna and in locating it in its proper position on the ship, no further precautions need be taken with the antenna and its associated feed system, in ordinary work. That is, as far as field patterns are concerned it is not necessary to scale every detail of the feed system nor to be sure that the antenna is matched to the detecting system. An exception to this general rule is met in the case of multiple-antenna systems, in which two or more individual antennas are fed with currents of definite relative magnitude in definite phase relationship. In such cases it may be necessary to scale every detail of the antenna system exactly and to be sure that impedances are matched throughout the system.

Although the conditions upon which the principle of similitude is based are not completely fulfilled in that no attempt is made to scale either conductivity or dielectric constant, there is ample experimental evidence that this defect introduces negligible error.

18.4.3 Calculation of Aircraft Antenna Patterns

In the v-h-f range particularly, the radiating system formed by the antenna and the skin of the ship upon which it is mounted may be exceedingly complex. Here the dimensions of structural parts of the plane are of the same order of magnitude as, or large compared to, the operating λ; such structural members tend to become increasingly effective in casting shadows, in causing reflection effects, and in generally disturbing the resultant field pattern as the frequency increases. Because of the similarity in size between such parts of the ship's structure and λ it is possible that resonance effects will occur in tail fins, stabilizers, guns, and in other antennas. It is further possible that resonance effects may occur in the smooth unobstructed skin of the fuselage itself; such resonant surface currents may have radiation characteristics that entirely mask that of the antenna proper, the antenna functioning some-

what like a coupling loop by means of which the skin of the ship is loaded. Because of all these possibilities and because of the geometrical complexity of the surfaces of modern aircraft, it is usually impossible to calculate the field pattern of a v-h-f aircraft antenna with any degree of accuracy.

At very high frequencies and under certain conditions it is sometimes possible to make pattern calculations that are qualitatively useful. These possibilities are limited to cases in which the antenna is located remote from reflecting and resonating objects, on or near smooth clean surfaces of extent large compared to the operating λ. As an example of a pattern calculation under such conditions, consider Figure 6. Here the fin constitute semi-infinite conducting planes intersecting at right angles. At 450 mc the surfaces involved are large compared to λ, and when the antenna is close in to the side of the fin the angle subtended at the antenna by the lower edge of that surface is large, approaching the 90° that would be subtended by the "edge" of a semi-infinite plane. It will be seen that measured and calculated patterns are in good agreement for the small spacing of $\lambda/2$ (corresponding to a subtended angle of 77°). At the larger spacing of 2λ (subtended angle 45°) the agreement is only qualitative in that both patterns have the same number of lobes in about the same position. At 4λ (subtended angle only 27°) the agreement is poor.

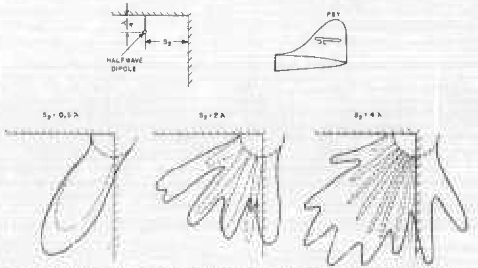


Figure 6. Discrepancy between calculated and measured field patterns of antenna under tail stabilizer of PBY. Dotted lines are calculated data; full lines as measured.

the solid lines represent the measured vertical-athwart-ship-plane patterns of a horizontal $\lambda/2$ dipole suspended with its axis in the line of flight $\lambda/4$ below the undersurface of the horizontal stabilizer of a PBY, at various distances out from the side of the vertical tail fin. The dotted lines represent the corresponding patterns calculated from simple image and antenna array theory, upon the assumption that the undersurface of the stabilizer and the side of

13.4.4 Definition of Polarization

Unlike the clear definitions of polarization used in optics, these concepts are used in aircraft radio engineering with considerable confusion. In the horizontal plane there is little difficulty, the electric vector of horizontally polarized radiation lying in the "horizontal" plane of the ship in normal flight; that of vertically polarized radiation lying normal to this

plane. In the vertical fore-and-aft plane the electric vector for horizontal polarization is always "horizontal" in that it lies parallel to the horizontal surfaces of the ship and is always in the same sense, that is, in the athwart-ship direction. In this plane the electric vector for vertical polarization is truly vertical only in two directions, dead ahead and dead aft. At all other angles of elevation in the vertical plane of flight "vertical" polarization has a horizontal component, proportional to the sine of the angle of elevation as measured from the horizontal plane. Directly above and directly below the ship "vertical" polarization is entirely horizontal in the usual geometrical sense, the direction of the vector being along the line of flight. In the vertical athwart-ship plane, "horizontal" polarization is always horizontal, the electric vector lying parallel to the horizontal surfaces of the ship, its direction along the line of flight; but "vertical" polarization is truly vertical only off the port and starboard wing tips, the electric vector being 100 per cent horizontal and athwart ship, directly above and directly below the ship.

There have been attempts in the past to avoid this confusion by the use of symbols, polarization being expressed in terms of its components along the two angular coordinates of a spherical reference system. This is correct and quite unambiguous to people having the spherical coordinate system perfectly in mind at all times, but such persons seem to be few and far between. Aircraft radio engineers continue to use the phrases "vertical polarization" and "horizontal polarization" with their customary promiscuity.

18.4.5 Presentation of Pattern Data

Many different methods of presenting pattern data in graphical form have been used in the past.

The most usual procedure is to give only three complete polar patterns for relative field strength or relative power in the horizontal, the vertical fore-and-aft, and the vertical athwart-ship planes. In most cases the distribution of radiation in these three planes suffices to give a pretty fair idea of the directivity of the antenna system.

Where complete information is desired, or where an unusual installation makes a peculiar field distribution probable, the complete spherical pattern, covering a solid angle of 4π, may be taken. Complete spherical data may be presented by means of three-dimensional models, by means of a series of plane polar diagrams, or by means of stereographic projection diagrams. An example of the latter form of presentation is shown in Figure 7. It has the big advantage over other methods of showing at a glance just where the radiation is going, with all lobes and nulls clearly evident, rapidly changing parts of the field being indicated by the crowding together of the constant-field-strength or constant-power contour lines. In the general case two such stereographic projections—one for each hemisphere—are necessary (for each polarization) for complete presentation of the data. But in the case of symmetrically located antennas a single diagram is sufficient for each polarization.

18.4.6 Absolute versus Relative Field Strength Patterns

It has been the practice in the past to present pattern data on a relative basis, simply in terms of arbitrary field strength or power units plotted against azimuthal or elevation angle. There has been some interest in the presentation of aircraft antenna pattern data on an absolute basis, in terms of millivolts per meter watt input power per mile, or in terms of the directivity of the system with respect to an isotropic radiator, a Hertz doublet, or a $\lambda/2$ dipole.[a]

Model measurements *can* be made to yield absolute patterns, but the procedure involved is extremely tedious.

[a] If D is the power directivity referred to an isotropic radiator, then:

$2/3D$	is power directivity with respect to a Hertz doublet
$\sqrt{2/3D}$	is field-strength directivity with respect to a Hertz doublet.
$0.61D$	is power directivity with respect to a $\lambda/2$ dipole.
$0.78\sqrt{D}$	is a field-strength directivity with respect to a $\lambda/2$ dipole.
$3.40\sqrt{D}$	is absolute field strength in millivolts per meter per watt per mile.

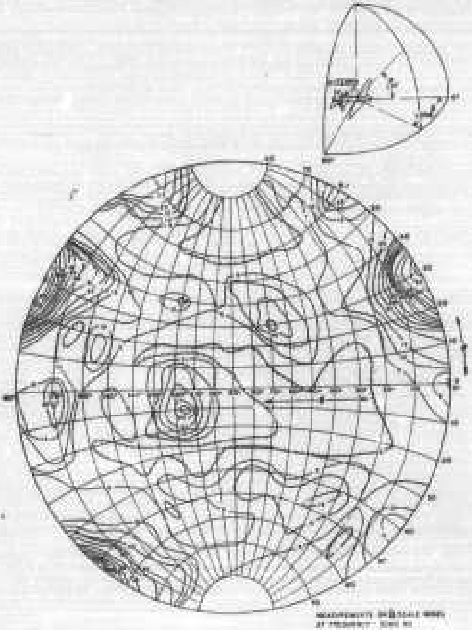


FIGURE [illegible]. Stereographic projection field pattern, [illegible] antenna on tail; power distribution in standard hemisphere.

18.7 Similarities between Aircraft Antenna and Ideal Antenna Patterns

The radiation maximum in the vertical pattern of a $\lambda/4$ (or less than $\lambda/4$) antenna worked against flat infinite ground is along the horizon, the entire field pattern of such a system being simply the upper half of the pattern of a vertical dipole in free space. This is not the case with ground planes of finite size.

Carter[a] has measured the vertical plane patterns of $\lambda/4$ antennas mounted on circular and rectangular ground planes of various size, and has found an approximately linear relation between the angle of throw-up of maximum radiation and the logarithm of the distance in λ from the base of the antenna to the edge of the surface upon which the antenna is mounted, the angle of throw-up being less the larger that distance, the theoretical value of zero elevation for an infinite plane being approached very slowly as the size of the surface becomes large in terms of λ.

This same effect occurs in the case of stub and whip antennas mounted vertically on aircraft. Figure 8 shows a plot of throw-up angle as a function of relative distance from the antenna was mounted on a smooth unobstructed logarithmic scale, the linear relation obtained by Carter for simple flat ground screens being indicated by the straight line. The twelve points shown in this figure were obtained from measured vertical plane patterns of stub antennas on aircraft in installations such that the antenna was mounted on a smooth unobstructed surface of curvature small compared to the operating λ in the direction in which the patterns were run. It is evident from the close agreement between these two sets of data that many antenna installations are found in practice under conditions such that the surface on which the antenna is mounted is a close enough approach to a flat ground plane so that Carter's data may be used to advantage in predicting the general n[illegible]ture of the distribution of radiation in the vertical planes. This simple relation

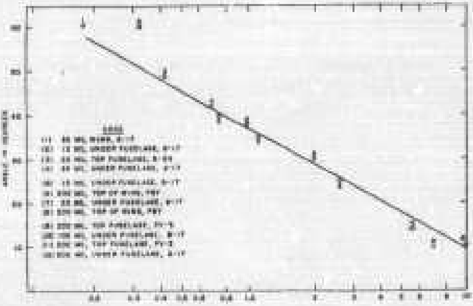


FIGURE 8. Angle of maximum radiation as function of distance from antenna to edge of surface on which antenna is mounted. Solid line represents experimental data for $\lambda/4$ antennas on flat ground planes.

will be quite invalid for antennas mounted on surfaces having pronounced curvature near the antenna, or for installations such that obstructions like fins, motors, turrets, and other antennas are likely to effect the field pattern in the plane in question.

A similar correspondence between aircraft antenna patterns and the patterns of antennas in ideal locations is found in the case of v-h-f and u-h-f horizontal antennas mounted near flat, horizontal surfaces on aircraft. Patterns quite similar to theoretical patterns are actually observed in the case of horizontal antennas mounted near a wing or fuselage surface that is large in terms of wavelength. In the case of aircraft installations, the field strength will not drop off to zero along the horizon because of the finite size of the aircraft surfaces.

18.4.4 Field Patterns of Vertical Antennas on Aircraft

Pattern versus Frequency

The field patterns of a given type of antenna in a given location on a given plane are markedly sensitive to frequency. Patterns of a given antenna on a given plane (see Chapter 17) reveal the following frequency-dependent phenomena:

In the Horizontal Plane. At the lower frequencies the horizontal plane patterns are fairly symmetrical, the vertical members of the ship's structure (for example the vertical fin of a B-17) being too small relative to the operating λ to be capable of casting sharp shadows or of causing pronounced reflection effects. At the higher frequencies these disturbing structures become large relative to λ, and the horizontal patterns are then more complex. Definite shadow regions appear, in which the field intensity is small compared to the average, although rarely zero owing to diffraction around the edges of the obstacles. Other minima are doubtless due to destructive interference between the direct ray in certain directions and the ray reflected from the tail-fin surfaces. Similarly maxima also appear in the pattern, resulting from constructive interference between direct and reflected rays.

In the Vertical Fore-and-Aft Plane. Even though the antennas (of a B-17 for example) may be mounted atop the fuselage over the wings, a great deal of radiation is found below the horizon at the low frequencies, at which the wing and fuselage surfaces are too small in terms of λ to act as efficient screens. At the higher frequencies little radiation is found below the horizontal plane.

At the lower frequencies the angle of maximum radiation is higher than at higher frequencies, agreeing with measurement and theory in the case of stub antennas worked against ground planes of finite size, showing that the smaller the extent of the ground plane in terms of λ the greater the angle of throw-up of maximum radiation.

Another effect evident in vertical fore-and-aft patterns is the increasing number of lobes and nulls as the frequency is increased. These are partially explained on the basis of structure effects, which would naturally be more pronounced the higher the frequency, but would also appear if the surface of the plane were absolutely flat and unobstructed, due to the presence of standing surface waves on ground planes of finite size.

In the Vertical Athwart-Ship Plane. The field patterns in this plane may be marked by similar effects, such as decreasing radiation on the opposite side of the ship from that on which the antenna is located, decreasing angle of throw-up of maximum radiation, and decreasing symmetry of pattern, as the frequency is increased.

Pattern versus Location of Antenna

The field pattern of a given type of antenna for a given frequency on a given plane, is greatly dependent upon the location of the antenna on the plane. This dependence is best demonstrated by reference to experimental results.

Figures 9 and 10 show the field patterns of a 100-mc λ/4 stub antenna in four widely different locations on a PBY. The great effect of location is obvious in this series. The patterns for the installation atop the vertical stabilizer are of particular interest in that the effectiveness

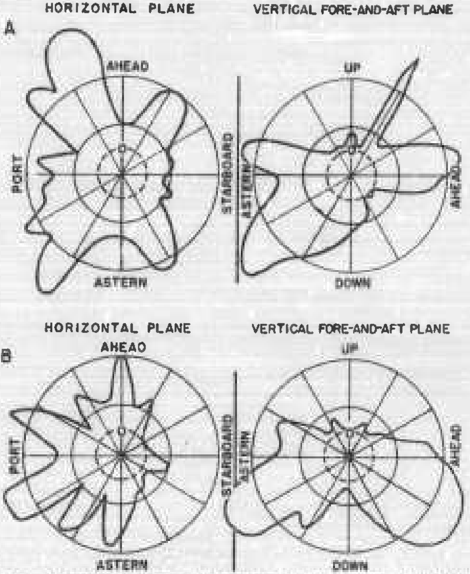
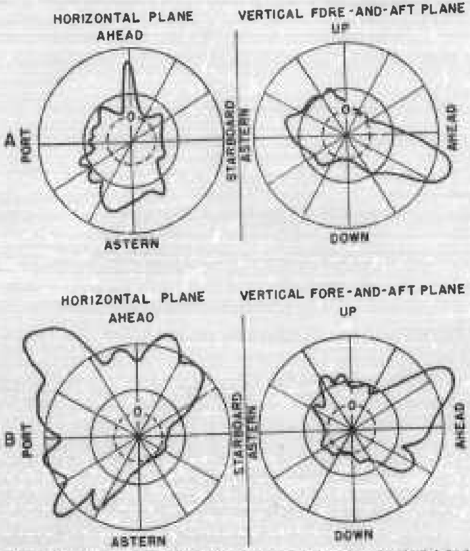


Figure [illegible]. Antenna pattern versus location, 1/4 stub at [illegible] mc located as follows: A, on top of vertical stabilizer; B, under port wing, [illegible] cm out from centerline of ship.

of a large surface (the wings) as a reflector (notice the large forward lobe in the upper vertical fore-and-aft pattern) and the ineffectiveness of a small surface (the tail stabilizer) as a screen (note the relatively large amount of downward radiation in the same pattern) are both demonstrated. The horizontal plane patterns of this series are interesting in that they



[illegible]

show the effect of slight deviations of antenna location from the axis of symmetry of the plane upon the symmetry of the horizontal pattern.

Another unusual example of the effect of location on field pattern is shown in Figures 11 and 12. The project in connection with which these patterns were taken involved a series of antennas for a B-24, the antennas to yield "uniform" distribution of vertically polarized radiation in the horizontal plane and in the upper hemisphere for 30° to 40° above the horizon. At 20 to 40 mc, suitable patterns could be real-

ized by means of antennas installed atop the fuselage, on the line-of-flight centerline, approximately 12 ft forward of the leading edge of the horizontal stabilizer. To simplify installation problems it was desired to mount the high-frequency antennas (40 to 60 mc) in this same location. While the horizontal and vertical fore-and-aft plane patterns are satisfactory, Figure 11 shows that the vertical athwart-ship pattern indicates that most of the radiation in that

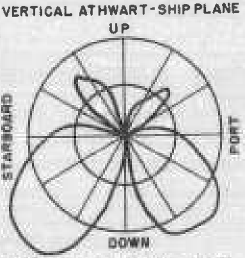


FIGURE 11. Effect of loading 40- to 60-mc antenna in same location as an antenna for [illegible] mc. Sleeve antenna, overall length [illegible], sleeve length [illegible]; 48 mc; vertical polarization; location vertical atop fuselage on centerline 11.5 ft forward of leading edge of horizontal stabilizer on [illegible].

plane is downward, and that there are sharp nulls about 20° above the horizon to either side of the ship. For these reasons this location for the h-f antennas had to be abandoned, and the 48-mc antenna was moved forward along the fuselage centerline to the trailing edge of the wing. Now the patterns in all three planes were satisfactory, the radiation in the vertical athwart-ship plane (Figure 12) now being uniform and upward as desired. This effect, believed due to the loading of the surface of the cylindrical fuselage when its circumference is resonant, is discussed in greater detail in the treatment of the broad-band whip antennas elsewhere in this report. In the case of this particular problem, such loading effects with consequent strong radiation from surface currents in the skin of the fuselage were inhibited by moving the antenna forward to where the wing surfaces could interfere.

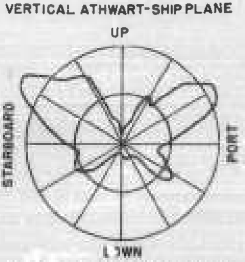


FIGURE 12. Moving 48-mc antenna of Figure 11 forward along fuselage centerline to trailing edge of wing produced desired upward radiation instead of undesired downward radiation as in Figure 11.

PATTERN VERSUS AIRPLANE

The pattern of a given type of antenna for a given frequency is greatly dependent upon the type of aircraft upon which it is mounted. This is particularly true in the middle and upper v-h-f range. The effect of the nature of the plane is most pronounced when the antenna is worked at frequencies such that sizes of structural members of the ship are of the same order of magnitude as the λ, and when the antenna is located on surfaces of pronounced curvature. About all one can say as to the patterns of similar antennas in similar locations on different planes, is that if the planes are much alike, differing mainly in size, then the pattern characteristics found on one plane will be found on the other plane, at a higher or lower frequency, depending upon the relative sizes. In the gen-

eral case, where v-h-f and u-h-f antennas are mounted on planes of widely different dimensions and shapes, the patterns will be greatly dependent upon the nature of the plane and upon the relation of the antenna to the predominate structural features.

18.4.9 Effect of Nearby Structures on Patterns

One of the hazards of aircraft antenna design is the possibility that an antenna designed to have certain pattern characteristics and installed on the ship in question will later be ruined by the installation of another antenna, a new turret, or an auxiliary gas tank in the immediate vicinity of the first antenna. Unless the effects of such disturbing structures are allowed for in the design of the original antenna, it is well to locate any metallic object of size or length comparable with the antenna dimensions at least one wavelength from the antenna.

18.4.10 Cross Polarization from Aircraft Antennas

There is usually some of the opposite polarization present in the field pattern of a simple vertical or horizontal antenna on aircraft, particularly in the lower v-h-f range. While the actual percentage of energy in cross polarization for a given installation depends in a complicated manner upon the frequency, the size and shape of the ship, and the location and orientation of the antenna, it is fairly safe to say that under ordinary circumstances it is small compared to the normally polarized radiation.

Under certain conditions, when the currents in the skin of the plane are properly disposed, the amount of cross polarization may be much greater. Also, the amount of cross polarization may be greater in other planes than the horizontal. While the presence of relatively large amounts of horizontal polarization is found with vertical antennas in particular installations, the phenomenon is not one upon which one may count in general; that is to say, it is ill-advised to expect to receive or transmit horizontally polarized radiation efficiently with a vertical antenna.

Cross polarization is usually more pronounced with horizontal antennas than with vertical, particularly in the h-f and lower v-h-f ranges. Here two factors are at work: (1) part of the vertical radiation may be due to loading of the skin of the ship or to resonance effects in vertical surfaces of the structure of the plane; and (2) at very low frequencies a horizontal antenna may have an appreciable vertical component, which, although small compared to the total antenna length, may be a relatively much more efficient radiator than the horizontal component, especially since the horizontal antenna current tends to be inhibited in its radiational effects by the presence of an opposite image current in the nearby surface of the ship.

18.4.11 Conclusion

Most of the pattern problems discussed above are serious only in the v-h-f range. At higher frequencies the pattern of a given antenna worked against a large clean surface on aircraft, will be quite similiar to the pattern on an infinite ground plane, except, of course, in cases where obstructions exist in the near field of the antenna. The effect of disturbing factors is then more pronounced the higher the frequency.

18.5 ANTENNAS FOR VERTICAL POLARIZATION

Compared to the problem of obtaining good aircraft antennas for horizontal polarization the design of vertical antennas for the v-h-f and u-h-f ranges is relatively easy. Not only is it easier to secure good input-impedance characteristics consistent with satisfactory mechanical and aerodynamical features, but, because of the symmetry inherent in the field of simple vertical antennas and because of the wide variety of locations in which they may be mounted on a plane, the problem of obtaining satisfactory radiational characteristics on aircraft is also much less difficult.

18.5.1 Broad-Band Antennas

The design of antennas having flat impedance characteristics is easier the higher the frequency. There are two reason for this; (1) the higher the frequency the smaller the physical dimensions of an antenna structure large in terms of λ and the less difficult the problems of mechanical and aerodynamical design, and (2) the higher the frequency, the smaller the antenna, and the less difficult the problem of feeding the antenna structure in a manner which will not impair its inherent broad-band characteristics. For these reasons antennas having band widths of the order of several octaves are practical in the u-h-f range, while in the lower v-h-f range it is a triumph of design to obtain a flyable antenna having a band width of only 20 to 30 per cent.[d] Bearing these facts in mind we proceed to a discussion of several successful broad-band antenna designs.

Wide-Band U-H-F Cone Antennas

The flat impedance characteristics of biconical antennas fed by balanced two-wire transmission lines were demonstrated experimentally and theoretically by Carter, by Schelkunoff, and by others, years ago.[3] More recently the Radio Research Laboratory has developed single unbalanced cone antennas for broad-band use on aircraft at ultra-high frequencies.

The cone antenna consists of a sheet-metal circular cone tapering from a small diameter at its base, where it is attached to the inner conductor of the coaxial feed line or to the inner conductor of a coaxial taper leading into the feed line, outward to a diameter of the same order of magnitude as its height. The apex angle varies for different applications, a typical value being 60°. The top of the cone is capped, either with another section of a cone of greater apex angle or with a segment of a spherical surface. The impedance characteristics of the cone antenna are remarkably flat over a range of frequencies corresponding to a range of antenna-height-to-wavelength ratios extending from about 0.2 to 2 or more. For many applications the cone antennas are sufficiently broad band in themselves to permit their being fed directly from the feed line without need for conventional matching sections. Figure 13 shows a sketch of a cone capable of covering the entire u-h-f range, 300 to 3,000 mc, with less than 2.5:1 *SWR* on a 50-ohm feed line.

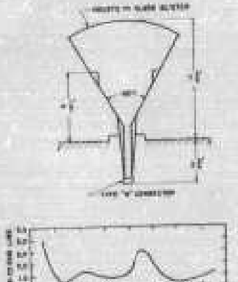


FIGURE 13. Wide-band cone antenna covering range 300 to 3,000 mc. (From RRL Report 411-TM-79.)

Cone antennas may be supported by insulating brackets attached to their peaks, or may be mounted within lucite radomes or blisters. Because of the large cross-sectional dimensions of the cone, or its surrounding blister, these antennas are not suitable for use on aircraft at frequencies much lower than 300 mc.

The patterns of cone antennas are in general similar to those of simple cylindrical radiators of equal electrical length. If desired, the cone antennas may be mounted at an angle with the

[d] Percentage band width = $(F_{max}/F_{min} - 1) \times 100\%$ where F_{max} and F_{min} are the upper and lower limits to the frequency range over which the antenna is matched to some specified standard, usually to a better than 2:1 *SWR* on a 50-ohm line.

side of the ship or in a horizontal position, in order to secure various amounts of horizontally polarized radiation.

The Sleeve Antenna (for Upper V-H-F and Lower U-H-F)

The sleeve antenna, developed by RCA Laboratories, is essentially a $\lambda/4$ stub surrounded for about half its length by a coaxial sleeve which may be simply an extension to the feed line. The sleeve is grounded to the skin of the ship at its base, the antenna being fed at the mouth of the sleeve. Since this feed point is approximately half-way up from the base, in a low-current region, the apparent input resistance is high, of the order of 100 ohms. This constitutes a big advantage in broad-banding, since a high-impedance antenna can be matched *down* to a 50-ohm line over a much wider band than that over which a low-impedance antenna can be matched *up*, other things being equal. The sleeve antenna differs from simpler broad-band antennas in that the attainment of flat input impedance characteristics is not of primary importance. The sleeve antenna involves four adjustable parameters—the ratios of sleeve length/total length and sleeve diameter/stub diameter in addition to the basic parameters (i.e., length/wavelength and diameter/length) of the simple stub—which make possible a high degree of control over the impedance characteristics of the antenna. By properly manipulating these four variables one can attain characteristics which are not necessarily flat and which may vary rapidly with frequency but which vary *in the right way* to "track" with a preselected type of matching section. Because of this flexibility, the sleeve antenna can be used to take better advantage of the properties of a simple matching section than can be obtained with antennas affording less control over input impedance. The effectiveness of this approach to broad-banding, i.e., that of distorting the antenna impedance to fit the characteristics of a given matching section, can be striking, a tenfold increase in useful band width resulting from the application of a properly designed matching transformer to a properly designed sleeve antenna being not unusual. A sketch of a simple sleeve antenna, and a typical *SWR*-frequency curve are shown in Figure 14.

Since the current distribution on the outer surfaces of the sleeve antenna is very much the same as that existing on the surface of a simple stub, the field patterns of sleeve antennas on aircraft will be similar to those of whip or stub antennas in corresponding locations.

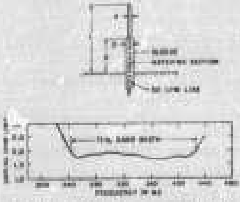


FIGURE 14. Schematic of sleeve antenna and SWR for given dimensions. [illegible]

The sleeve antenna is most useful on aircraft in the general range 100 to 1,000 mc. At higher frequencies a cone antenna may be used to better advantage, while at lower frequencies the cross section of wide-band sleeves becomes prohibitively large, so that broad-band whip antennas are more satisfactory.

Broad-Band Whip Antennas

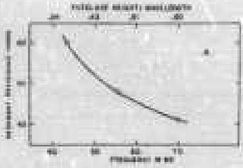
Broad-band whip antennas developed by the Antenna Section Research Division, Aircraft Radio Laboratory, Wright Field, constitute some of the most successful aircraft antennas. In addition to their electrical features these antennas have very low wind drag, are tactically inconspicuous, are easily mass-produced, and are relatively easy to install on aircraft.

The antennas are tough steel whips, the diameter tapering from about ½ in. at the base

to about ⅛ in. at the tip, the height (roughly $\lambda/4$) ranging from 30 in. to 6 ft, depending on the frequency band. When used in conjunction with a simple two-element matching section consisting of lengths of standard coaxial cable compactly coiled in a metal can attached to the base of the whip, these antennas are capable of approximately 40 per cent band width in the lower v-h-f range. Several such antennas, each with its associated matching section, have been designed to be easily interchangeable in a single mounting fixture, four of them together covering the 36- to 110-mc band with less than 2:1 *SWR* on 50-ohm cable.

The whip antennas depend for their broad-band characteristics partly upon the fact that the impedance level of a stub antenna worked against a cylindrical ground surface having a circumference of the order of magnitude of the operating λ is higher than that of the same antenna worked against a flat ground plane. This fact is shown by the curves of Figure 16, which represent the variation with frequency of the resonant resistance of stub antennas identical except for length mounted on curved surfaces. Both curves indicate that impedance levels much higher than the normal 36 ohms of a stub antenna worked against flat ground can be obtained with stub or whip antennas mounted on the roughly cylindrical fuselages of large planes in the lower v-h-f range.

Evidence indicates the necessity for basing the design of antennas for the lower v-h-f range upon impedance measurements made with the antenna installed in the location in which it is to be used. While such measurements are preferably made on the actual ship, or on a partial full-scale mock-up it is possible, if great care is taken in scaling both the plane and the antenna and its feed system, to obtain useful results with measurements made on models. While the effect of the fuselage upon the input characteristics of broad-band whip antennas is favorable, both as regards high-impedance level and flat reactance characteristics, if a whip antenna incorporating a matching section based upon flat-ground-plane impedance measurements were used in the same location as these whips, the results would not be satisfactory. The characteristics of the antenna are too greatly affected by currents in the curved fuselage for ground-plane measurements to be valid. This effect is not limited to the lower v-h-f range, but occurs in any aircraft antenna installation where the radius of curvature of the skin of the ship at the antenna location is small compared with the operating λ.



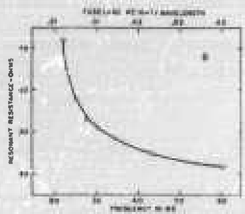


FIGURE 16. Characteristics of A, similar whip atop fuselage of B-24, 15 ft aft of trailing edge of wing, full-scale ship in flight (data from ARL); B, stub antennas below fuselage of B-24 on center-line of wing and fuselage, ⅛-scale model (data from RCA Laboratories).

Figure 16 shows a typical *SWR*-frequency curve for a low-frequency broad-band whip mounted on a B-24. The band width shown, obtained by means of a simple two-element transmission line matching section, is more than twice as great as that obtainable with the

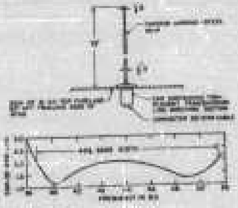


FIGURE 16. Broad-band whip antenna for B-24. (Data from ARL Report 360.)

same antenna mounted on a large flat ground plane and used in conjunction with a much more complicated matching section.

Figure 17 shows the principal plane patterns of whip antennas mounted atop the fuselage of a B-24. The typical butterfly distribution evident in the vertical athwart-ship planes is in fair agreement with that calculated by Carter for stub antennas worked against cylindrical surfaces of corresponding relative size. The effect of currents in the skin of the fuselage can be seen by comparing the vertical athwart-ship plane patterns with those for the vertical fore-and-aft planes; the former show much more radiation in directions below the horizon than do the latter, as would be expected since the top of the fuselage represents a much closer approach to large flat ground plane along the line of flight than it does athwart ship.

Broad-Band Fan Antennas

A fan-shaped array of three or more wires, strung from a lead-in on the side or top of the fuselage to some supporting structure, such as a guy wire or part of the ship itself, forms a satisfactory broad-band antenna at the extremely low-frequency end of the v-h-f band. These antennas, developed by the Antenna Section, Research Division, Aircraft Radio Laboratory, Wright Field, have numerous advantages.

1. They have impedance characteristics such that they are easily matched to 50-ohm cable over frequency bands of the order of 25 to 35 per cent wide, an unusual band width in view of the low frequencies involved. Impedance matching of fan antennas is usually effected by means of a two-element matching section consisting of lengths of commercially available coaxial cable compactly coiled in a metal container mounted just after the lead-in, inside the ship

2. Since the fans are made of ordinary aircraft-antenna wire they offer much less wind resistance than do conventional large-surface-area antennas of comparable band width at comparable frequencies.

3. Because of the fineness of these wires a

FIGURE 17. Pattern of whip antenna installed on B-24, top centerline 55 in. behind trailing edge of wing, 40 mc.

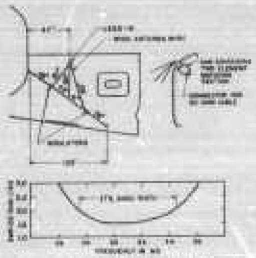


FIGURE 18. Broad-band fan antenna for B-24. (Data from ARL Report 357.)

fan antenna is practically invisible at distances of the order of 20 ft or more, an obvious tactical advantage.

Among the disadvantages inherent in fan antennas, of negligible importance compared to their advantages for most applications, are the following:

1. Fans must be tailored especially for each installation. Because of their spread in area, the characteristics of fans are sensitive to minor differences in the structure of the plane in their immediate vicinity and to the presence of near-by antennas.

2. Installation of a fan is not a very convenient procedure.

3. The field patterns of fan antennas are not notably symmetrical, since to get sufficient height for a large percentage of vertically polarized radiation the antennas must usually be worked against one side or other of the ship. In some cases the asymmetry is such that a tactical course must be flown.

4. There is usually a considerable percentage of horizontally polarized radiation in the field of a fan antenna.

Figure 18 shows a sketch of a typical fan antenna installation and its *SWR*-frequency characteristic as measured in flight at Wright Field. Figures 19 and 20 show field patterns at the center of the band of this antenna, for vertical and for horizontal polarization, respectively, as measured by means of models by the Ohio State University Research Foundation.

BROAD-BAND INVERTED-L ANTENNA (LOW V-H-F)

The broad-band inverted-L antenna, developed by RCA Laboratories, is an adaptation of the simple inverted-L, or flat-top, antenna used on aircraft at frequencies so low that the

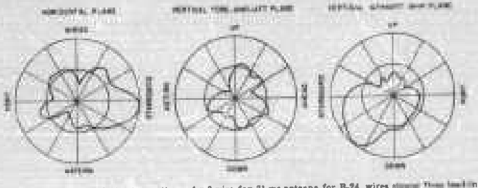
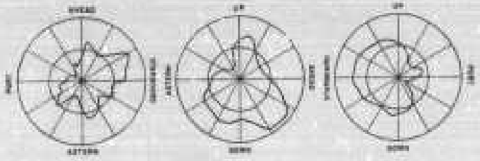


FIGURE 19. Vertical polarization patterns for 3-wire fan 31-mc antenna for B-24, wires strung from lead-in near top starboard fuselage 47 in. forward of leading edge of starboard stabilizer to point near bottom side of fuselage 105 in. forward of leading edge of vertical stabilizer. (From ARL Report 357.)



[illegible] from lead-in near top starboard fuselage, 47 in. forward of leading edge of starboard stabilizer, [illegible] bottom side of fuselage, 105 in. forward of leading edge of vertical stabilizer. (From ARL Report 357.)

height of a conventional $\lambda/4$ whip or stub would be prohibitively great. A sketch of a simple inverted L having a height equal to half its total length is shown in the upper portion of Figure 21, where the measured input impedance of such an antenna is also shown. It is evident from these impedance characteristics that while the antenna has a much smaller physical height at resonance than the corresponding whip antenna, its impedance level is low and its reactance characteristic is steep, limiting its usefulness to spot-frequency or narrow-band applications.

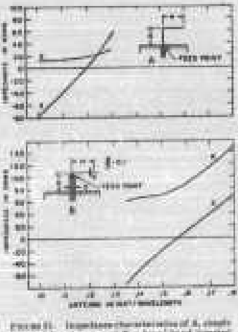


FIGURE 21. Impedance characteristics of A, simple inverted L antenna, and B, a broad-band inverted L antenna.

It has been found possible, by means of a sleeve (that is, by extending the coaxial feed line beyond the ground plane up to the bend in the antenna) and by properly proportioning the relative cross sections of the vertical and horizontal members of the antenna, to retain most of the reduction in vertical height gained in the simple inverted L and at the same time secure an antenna of broad-band characteristics. The modified inverted L is sketched in the lower part of Figure 21. The modified version has a much higher impedance level and an appreciably flatter reactance-frequency curve than the simpler antenna. Consequently it can be matched to a low-impedance line over much wider frequency bands. The actual band width obtainable with an inverted L depends upon the complexity of the matching section used; band width varies from about 12 per cent in the case of an L fed directly from 50-ohm line to about 58 per cent when the L is used in conjunction with a three-element matching section consisting of lengths of commercially available coaxial cable.

10.3.2 Quarter-Wave Antennas for Vertical Polarization

Among the more simple antennas suitable for use on aircraft in applications calling for limited band width in the u-h-f and upper v-h-f ranges are the following.

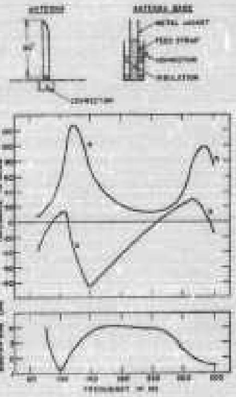


FIGURE 22. Thick stub antenna (AN-156). (Data from RRL Report 411-TM-86.)

THICK STUB ANTENNA

A $\lambda/4$ stub of fairly large cross section is known to have broad-band input characteristics. However, there are two difficulties inherent in these antennas: (1) The fact that they must be base-insulated requires the use of low-loss solid-dielectric mounting fixtures of great mechanical strength, and (2) the large base must be connected to the small inner conductor of a coaxial feed line in a manner which will not destroy the intrinsic broad-band characteristics of the antenna. This last is a difficult problem which has not been satisfactorily solved to date.

The AN-155 antenna, developed by the Radio Research Laboratory, is an example of the thick stub as used on aircraft. This antenna, sketched in Figure 22, consists of a phenolic-impregnated maple mast, covered, except at its base, by a metallic sheath. The base is held by an insulating bracket and the sheath is fed by a tapered metal strip, or "dog ear," connecting the lower edge of the sheath to the inner conductor of a standard coaxial cable connector.

The measured input impedance of this antenna is also shown in Figure 22. It is to be remarked that these characteristics depend to a rather large extent upon the shape and position of the "dog ear." Such a mast antenna, 30 in. high, and of 2¾x1¼ in. streamline cross section, is capable of covering the 90-to-110-mc band without need for external matching sections. By cutting down the length of the antenna, higher frequency bands, of increasingly greater width, may be covered.

BROWN-EPSTEIN ANTENNAS

A simple u-h-f antenna, combining a strong mechanical mounting with provision for moderate band width, is shown in Figure 23A. In its simple form the antenna consists of a $\lambda/2$ rod mounted coaxially in a $\lambda/4$ deep cylindrical well set into the ground plane against which the antenna is worked. The portion of the system contained in the well serves two purposes: (1) It acts as a shorted $\lambda/4$ line, which, by presenting a high impedance to ground at the central feed point, effectively insulates the base of the protruding $\lambda/4$ radiator while at the same time it provides a strong metallic mounting, and (2) this shorted $\lambda/4$ line acts as a parallel-resonant circuit in shunt with the antenna, a circuit well known to be effective in flattening the reactance characteristic and in raising the impedance level of a series-resonant antenna.

A version more suitable for use on aircraft is sketched in Figure 23B. Here the shorted-line support has been placed inside the radiator, where it functions exactly as before, with

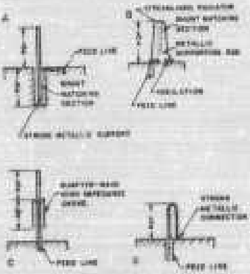


Figure 23. Antennas for vertical polarization: A, Brown-Epstein antenna; B, modified Brown-Epstein antenna; C, skin-back antenna; D, half-folded dipole.

the additional advantage that the radiating surface is now larger and consequently will have flatter impedance characteristics. This surface may be streamlined to reduce drag.

Skin-Back Antenna

The skin-back antenna, sketched in Figure 23C, consists of a $\lambda/4$ radiator which may be considered as the continuation of the inner conductor of the coaxial feed line, the outer conductor of which is folded back upon itself to form the lower half of a vertical $\lambda/2$ dipole. The shorted $\lambda/4$ line acts as a high-impedance choke in series with the lower half of the antenna, effectively isolating it from the remainder of the outer surface of the feed line.

This antenna has been used on aircraft in installations such that it is desirable to isolate the antenna from the adjacent surfaces of the ship. For example, it has been used atop the vertical fin of a B-17, where surface currents in the immediate neighborhood of the feed point of a conventional stub would greatly modify the stub impedance and radiation characteristics.

Half-Folded-Dipole Antenna

One-half a Carter folded dipole worked against ground (see Figure 23D) forms an aircraft antenna with two advantages over a simple stub. (1) It includes its own matching section, for, by proportioning the relative diameters of the two conductors properly, it is possible to secure a perfect match to the feed

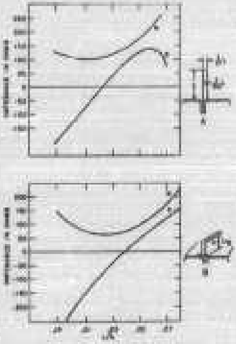


Figure 24. Impedance characteristics of: A, half-folded dipole, and B, two half-folded dipole.

line, and (2) since one side of the antenna is metallically grounded the antenna is inherently strong. Measured impedance characteristics of a typical half-folded dipole are shown in the upper part of Figure 24.

10.3.3 Less-than-Quarter-Wavelength Vertical Antennas

An obvious means of minimizing aerodynamic and mechanical difficulties while at the same time retaining some of the desirable features of the $\lambda/4$ antenna is to use resonant antennas which are short compared to the operating λ. There have been many attempts at antenna design along this line, four of which are shown in Figure 25.

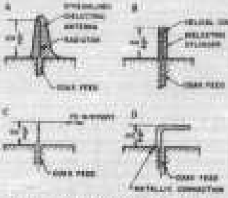


FIGURE 25. Antennas less than $\lambda/4$ long for vertical polarization: A, dielectric; B, helical; C, inverted L; D, bent half-folded dipole.

DIELECTRIC ANTENNA

A simple stub radiator, surrounded by dielectric material, will be resonant at a much lower frequency than the same antenna in air, the actual reduction in physical length depending upon the inductive capacity and relative volume of the dielectric. The impedance characteristics of such an antenna are plotted in Figure 26. It will be noted that the impedance level is too low to permit successful broad banding.

HELICAL ANTENNA

The helical antenna contains its own loading coil, and while it can be made to be resonant at a height equal to a small fraction of $\lambda/4$, its characteristics are marked by a very low-impedance level and by a very steep reactance characteristic so that its usefulness, if any, is limited to spot-frequency applications.

INVERTED-L ANTENNA

The inverted L has short physical height and, if modified, can be broad banded, as is described elsewhere in this report.

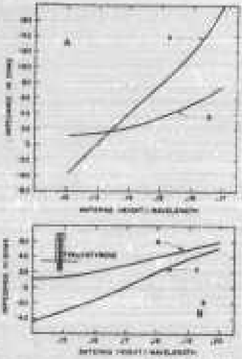


FIGURE 26. Impedance characteristics of A, helical antenna of Figure 25, and B, stub antenna in dielectric cylinder.

BENT HALF-FOLDED-DIPOLE ANTENNA

This antenna has impedance characteristics which are matchable to 50 ohms over moderate frequency bands.

10.3.4 Half-Wave Grounded-Loop Antenna

A semi-circular $\lambda/2$ loop antenna, mounted in a vertical plane, with one end grounded to the skin of the ship and the other end attached to the inner conductor of a coaxial feed line, has several advantages over the simple $\lambda/4$

stub for vertical polarization in the u-h-f range.

1. Its vertical height is less than one-sixth of the resonant λ, so that if mounted with the plane of the loop in the line of flight this antenna will have less wind drag than the corresponding stub.

2. Since one side of the loop is grounded it can be made to have great mechanical strength.

3. Its impedance characteristics are such that it can be easily matched over quite wide frequency bands.

The field pattern, for vertical polarization, of this antenna worked against a flat surface on aircraft will be similar to that of a vertical stub antenna in the same location. The patterns of full-wave loops have been investigated theoretically by Carter, and are discussed in Section 18.7.

18.6 MULTIPLE-RESONANT ANTENNAS

It may sometimes be desirable in aircraft antenna work to use a single antenna for transmission or reception in two or more separate frequency bands, the frequencies being so widely spaced as to make the use of a broadband antenna covering the entire range impractical. One scheme for realizing such an antenna system is shown in Figure 27. The antenna consists of a cylindrical stub made in two pieces, the upper portion being supported by the inner conductor of a shorted coaxial line which is recessed into the lower part of the stub. The lower part of the antenna is of such length as to be resonant at the center of the high-frequency band, the $\lambda/4$ line built into this section effectively isolating the lower stub from the upper portion of the antenna. The total length of the antenna is made such that, allowing for the reactance introduced by the line section, the complete antenna will be resonant in the center of the low-frequency band. It will be seen from the impedance curves of Figure 27 that the particular antenna shown is useful from 320 to 372 mc and from 468 to 520 mc, representing an extreme range of frequencies which could not be covered by a conventional antenna of like cross section. By varying the dimensions of the various parts of the antenna it is possible to secure pass bands of greater or less spacing and of greater or less band width. The principle can, of course, be extended to three or more pass bands. This antenna has one advantage

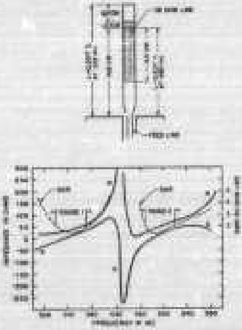


FIGURE 27. [illegible] having two resonant bands, [illegible] and [illegible] mc.

over very wide-band antennas, in that it avoids the pattern difficulties which may appear at the high-frequency end of the band of such antennas, where the antenna may be several multiples of $\lambda/4$ long.

18.7 ANTENNAS FOR HORIZONTAL POLARIZATION

The design of aircraft antennas for horizontal polarization in the v-h-f and u-h-f ranges generally involves much greater difficulties than the design of corresponding vertical antennas. Not only are there grave mechanical problems, particularly in the lower v-h-f band where the structure of the plane offers few alternative methods of supporting the antenna

and consequently permits only a restricted choice of antenna locations, but there are electrical problems as well. The pattern and impedance requirements for horizontal antennas at these frequencies are usually such as to demand that the two components of the antenna be fed in some definite phase relationship, usually 180° out of phase, with the result that balance transformers as well as conventional matching sections must be included in the feed system if band width is desired.

16.7.1 Broad-Band Balance Transformers

Since most antenna installations for horizontal polarization on aircraft are of the balanced type, and since most aircraft transmitters are designed to work into low-impedance

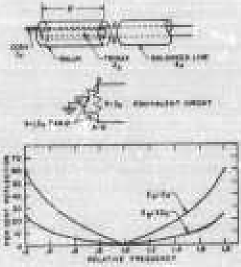


FIGURE 28. Broad-band balance transformer known as a "bazooka" or "balun." At bottom, reflection introduced by balun between [illegible] and matched [illegible] of impedance Z_0. (From RRL Report 411-TM-22.)

unbalanced cable, it is usually necessary to insert between the antenna and the line a device for maintaining balance even though the frequency departs considerably from resonance. One of many such transformers is the "bazooka" or "balun" developed by the Radio Research Laboratory, sketched in Figure 28. When properly constructed this transformer maintains both sides of the two-wire line at equal and opposite potentials with respect to ground over large frequency ranges to either side of that for which the twinax line inside the balun is $\lambda/4$ long. Furthermore, if the characteristic impedance of the twinax line in the balun is large compared to those of the coaxial and balanced cables between which it is inserted, the presence of the balun will cause little reflection over very wide frequency bands, as may be seen from the reflection-frequency curves of Figure 28. These curves apply to a rather artificial case, since the impedance of balanced cable or of balanced antennas is usually higher than that of coaxial cable. In practice a transforming section must usually be inserted on one side or other of the balun.

16.7.2 Antennas for "Uniform" Horizontal Plane Pattern

Bent-Sleeve Dipole Antenna

The sleeve dipole antennas developed by the Radio Research Laboratory give pear-shaped horizontal plane patterns for horizontal polarization and have broad-band characteristics at frequencies less than 600 mc. The antenna consists of a $\lambda/2$ dipole bent into a V having an included angle of about 100°; each arm of the V is surrounded for about half its length by a coaxial sleeve; the arms tie in to the balanced side of a broad-band balance transformer which in turn is fed by 50-ohm coaxial cable. The antenna and its attached balun form a unit which plugs into a streamlined cylindrical mount which is permanently attached to the skin of the plane, the mount holding the plane of the V horizontal, in proper relation to the skin of the ship and to the direction of flight. A series of such antennas may be used interchangeably in the same mount, in order to cover a very wide total frequency range.

Figure 29 shows plan and elevation sketches of this antenna, and includes a sketch of the feed system. The *SWR*-frequency curve of that figure gives some idea of the band width at-

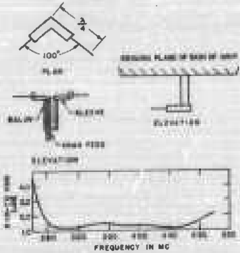


FIGURE 29. Plan and elevation views, feed system, and *SWR* of bent-sleeve dipole. (From RRL Report 411-TM-122.)

tainable in the upper v-h-f and lower u-h-f ranges. While these antennas have very satisfactory characteristics in the 200- to 600-mc range, their impedance characteristics are marred at higher frequencies by the adverse effects of the feed-system discontinuities on band width, and at lower frequencies they become physically large, introducing mechanical and wind-drag difficulties.

Figure 30 shows the pattern yielded by the bent-sleeve dipole mounted approximately $\lambda/4$ below the undersurface of the fuselage of a large plane. The large energy throw-down in the vertical plane could be remedied by moving the antenna farther out (closer to $\lambda/2$) from the skin of the ship.

THE COAXIAL-FED V-DIPOLE ANTENNA

In some respects this antenna is similar to the bent-sleeve dipole antenna. It consists of a coaxial-fed (unbalanced) dipole with $\lambda/4$ arms attached to the inner and the outer conductors of the supporting feed line, the arms lying in a plane perpendicular to the feed line and forming a V with an included angle of 95 to 100°. The supporting line extends $\lambda/4$ beyond the ground plane on which the antenna is mounted, its outer conductor being grounded at the base.

The horizontal plane pattern of the antenna is peanut-shaped with side minima in field strength about 4 db down from the maxima. The vertical plane patterns show the large amount of throw-down to be expected from a horizontal antenna mounted $\lambda/4$ from a conducting sheet.

The antenna yields little vertically polarized radiation at resonance, but as the length of the supporting line departs from $\lambda/4$, the currents in its outer surface result in increasingly larger percentages of vertical radiation. It has been found experimentally that the antenna can be used over frequency bands approximately 35 per cent wide before the maximum average per-

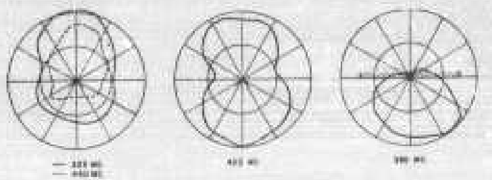


FIGURE 30. Patterns of bent-sleeve dipole for horizontal polarization. (From RRL Report 411-TM-122.)

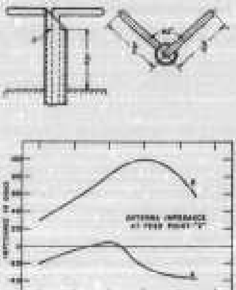


FIGURE 31. Coaxial-fed V dipole for horizontal polarization.

centage of vertical polarization becomes greater than 20 per cent of the total radiation in the plane containing the most vertically polarized energy. Since no effort is made to maintain balance between the two sides of the V, the patt ·ns tend to become asymmetrical at frequencies far from resonance, another factor limiting useful band width to about 35 per cent.

Figure 31 shows a sketch of a simplified version of the antenna, a typical set of input impedance characteristics, and a *SWR*-frequency curve for an antenna having a built-in low-impedance series transformer. Because of the simplicity of the feed system these antennas have wide-band u-h-f impedance characteristics, a set of four interchangeable antennas covering the 500- to 1500-mc range with less than 2:1 *SWR* on 50-ohm line.

Figure 32 shows the measured horizontal plane patterns at three frequencies distributed over the range of a model intended for use in the 1175- to 1500-mc band.

SPLIT-CAN ANTENNA

Figure 33 shows a sketch of a u-h-f split-can antenna developed by the Radio Research Laboratory for horizontal polarization. The antenna consists of a cylinder of streamlined cross section, split longitudinally along the trailing side, and mounted normal to a ground surface from which its base is insulated. The antenna is fed by a balanced line, which ties on at the two opposing edges of the split. In a tentative theory the edges of the split are regarded as a continuation of the two-wire feed line, the surface of the antenna acting as a shunt loop

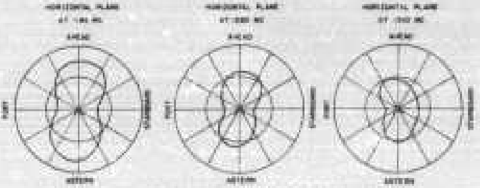


FIGURE 32. Measured field patterns for coaxial-fed V dipole, horizontal polarization.

across this line. Since the currents in this surface loop are horizontal, tho resulting radiation is horizontally polarized, the horizontal plane pattern being substantially uniform with the minimum only about 3 db down from the maximum. When the surface of the antenna is less than $\lambda/2$ around, its effect is that of an induc-

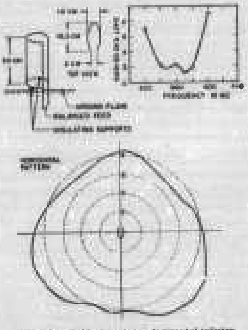


FIGURE 33. [illegible] antenna, horizontal polarization. (From [illegible] Report [illegible].)

tive shunt across the feed line, resulting in an increase in the physical length of the antenna at resonance. For example, the 30-cm high antenna of Figure 33 is resonant at approximately 375 mc, corresponding to an electrical length of 0.375 λ compared to 0.24 or less for a conventional resonant stub of similar cross section.

Loop Antennas

The possibilities of circular loop antennas of dimensions large compared with the operating λ have been explored, independently, by Carter,[4] Foster,[5] and Sherman.[6] Large loop antennas have characteristics quite different from those of the small loops used for direction finding at low frequencies, characteristics which make these antennas attractive in certain applications where horizontal polarization is desired in the u-h-f range.

In the absence of experimental data on the impedance and pattern characteristics of full circular loops on aircraft this discussion will be limited to the presentation of theoretical data taken from a report by Carter.[4] While the impedance level of loops less than $\lambda/2$ in circumference is low, loops of the order of λ in circumference have respectable input resistances. A loop $\lambda/2$ in circumference mounted with its plane horizontal would yield a very uniform pattern for horizontal polarization in the horizontal plane; unfortunately its radiation resistance would be only about 13 ohms, a value which would have to be stepped up considerably, possibly by means of a sleeve, before the antenna would be useful for any but narrow band applications.

12.7.3 Antennas of the $\lambda/2$ Dipole Type

Antennas of the center-fed $\lambda/2$ dipole type are commonly used for horizontal polarization in applications such that the nulls in the field pattern along the direction of the antenna axis are not objectionable.

Brown's Antenna

An interesting u-h-f antenna for horizontal polarization on aircraft is that developed by G. H. Brown. The $\lambda/4$ arms of the dipole consist of strong tubing held in line by an axial insulating rod, the dipole being supported in a horizontal position $\lambda/4$ out from the skin of the ship by means of two vertical metal cylinders, closely spaced and connected to the two arms of the dipole at either side of the central feed point. Since the vertical supports are base-grounded, a mechanically strong mounting is secured, while the fact that these supports are $\lambda/4$ long insures that the antenna is electrically insulated from ground. A coaxial feed line, which may include a series transformer matching section, runs up through one of the sup-

porting cylinders, tying on to the two sides of the dipole at the feed point.

In another version of this antenna the axial rod aligning the two halves of the dipole is metallic and connected to the radiators only through metal plugs at each end of the dipole. This system is fed from a balanced line, one side of the line tying on to each radiator at the feed point, the coaxial lines included inside the radiators acting as a shunt matching section, which results in flat input impedance characteristics over a fairly wide frequency band.

Wire Dipoles

Because of mechanical and wind drag considerations none of the antennas previously considered are suitable for use on aircraft at frequencies much below 200 mc. For frequencies in the lower v-h-f range it becomes necessary to use wire antennas, an example being a $\lambda/2$ dipole of ordinary aircraft antenna wire strung parallel to the line of flight, supported at its two ends either by masts or by convenient points of the ship's structure, and center-fed from a twisted pair or twinax cable. Such antennas are, of course, extremely narrow band, and are therefore useful only for spot-frequency applications, or for applications in which manual or mechanical antenna tuning is permissible.

The band width of wire dipoles can be considerably improved by making each radiator in the form of a cylindrical or conical cage of wires, thus simulating large-surface conductors while at the same time retaining the low-drag features of wire antennas. The average characteristic impedance of multi-wire cage dipoles is discussed at length in a report prepared by Division 15.[7]

18.7.4 Polyphase Antennas

Another example of antenna system which has u-h-f possibilities, and perhaps even for much lower frequencies, is the turnstile antenna developed by Brown and by Lindenblad for f-m and television transmitting purposes. This antenna, which may be regarded either as two crossed $\lambda/2$ dipoles fed in phase-quadrature or as four $\lambda/4$ antennas arranged along the diagonals of a square and fed in 90° phase rotation, yields an unusually symmetrical pattern for horizontal polarization in the horizontal plane, and has the further advantage of naturally broad-band characteristics in that the reflection coefficient on the main feed line is equal to the square of that existing on the branch lines leading to the individual antennas. The desirable features of the turnstile antenna are accentuated in the three-phase Y antenna, which, since it has only three radiators, instead of four, is perhaps more attractive for low-frequency use on aircraft. This antenna consists of three $\lambda/4$ radiators arranged symmetrically in the horizontal plane, the radiators being fed with equal currents in three-phase relationship. In this system the reflection coefficient on the main feeder is equal to the cube of that existing on the individual branch lines, resulting in still greater broad banding due to feed than is obtained with the turnstile antenna.

Despite their advantages there are very good reasons why polyphase antenna systems have not been exploited thus far. In the first place these systems require high-impedance component antennas, of 100 ohms input impedance in the case of the turnstile, and 150 ohms in the case of the Y, presuming a 50-ohm main line. While high-input-impedance-$\lambda/4$ antennas may be obtained by means of sleeves there remains the problem of obtaining high-impedance branch lines. Furthermore the patterns of these antennas depend upon proper phasing of the currents in the component antennas, which in turn depends upon a proper impedance match throughout the system. For this reason it is difficult to measure the field patterns of these antennas on aircraft by means of models, since not only must the antenna dimensions be properly scaled, but the individual antennas must be accurately matched to their feed lines. The latter condition is difficult to realize at the s-h-f range used in model work. Whether the large amount of experimental work required in the development of these antennas is justified depends upon the need for uniform pattern and broad-band impedance characteristics. The fact remains that these are among the very few antennas that have even a chance of satisfying

such requirements on aircraft in the lower v-h-f range.

18.7.5 Multiple Antennas for Horizontal Polarization

Any of the antennas suitable for vertical polarization may be used for horizontal polarization if properly mounted. However, such antennas, mounted on one side of the fuselage in a horizontal position or at an angle with the side of the fuselage, necessarily give asymmetrical patterns. If symmetry is desired it is necessary to use two similar antennas mounted in corresponding positions on opposite sides of the plane and fed in proper phase relationship. Such antenna combinations have been developed by the Radio Research Laboratory in collaboration with the Ohio State University Research Foundation. When antennas made up of $\lambda/4$ stubs and cones are fed in phase, the horizontal plane pattern contains lobes ahead and astern for vertical polarization, with nulls in the corresponding positions for horizontal polarization; when the antennas are fed 180° out of phase the situation is reversed. As far as symmetry is concerned, these antenna systems become less satisfactory the higher the frequency.

Since no compensation is gained in in-phase or in out-of-phase feeding, individually broadband antennas must be used if band width is desired.

18.7.6 Surface or Interior Antennas

Flush-mounted antennas such as slots, horns, and wedges are suitable for horizontal polarization at upper v-h-f and u-h-f frequencies. In many cases, such antennas, mounted singly on the underside or wing or fuselage or in pairs on either side of the ship, have pattern and impedance characteristics satisfactory for certain applications. These antennas are described elsewhere in this report.

18.8 ANTENNAS FOR BOTH VERTICAL AND HORIZONTAL POLARIZATION

While any of the conventional linear antennas can be mounted at odd angles with the skin of the ship in order to secure varying amounts of both horizontal and vertical polarization, and while other antennas, such as fans and loops, incidentally give radiation of both types, there are special antennas for this purpose.

The Fish-Hook Antenna for Circular Polarization

The M2201 and M2202 antennas developed by the Radio Research Laboratory consist of two thick dipoles, crossed at right angles, with the individual radiating elements bent downward at an angle of approximately 30° with the horizontal. Each radiator is supported by one conductor of a four-conductor open line $\lambda/4$ long. This line leads into the interior of the ship to a phasing and matching unit which feeds the two dipoles in phase quadrature and which matches the combined input impedance to the main transmission line. A sketch of the antenna, together with its *SWR*-frequency characteristic and measured field patterns, is shown in Figure 34.

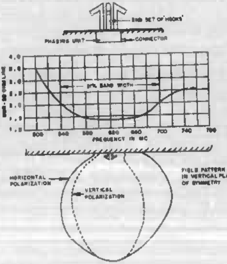


Figure 34. Fish-hook antenna for circular polarization. (Data from RRL Report 411-TM-58.)

The antenna is intended to be mounted on the underside of the fuselage of a plane, the maximum radiation being dowoward with ap-

proximately equal vertically and horizontally polarized components. The antenna feed system is so arranged that it can be used with either one or two transmitters. It is designed to be mounted inside a plastic radome.

Trailing-Wire Antennas (H-F and Low V-H-F)

The simple trailing-wire antenna used on aircraft for long-range communication in the m-f and h-f ranges is an example of an antenna with which greater or less amounts of either vertical or horizontal polarization may be obtained. It consists simply of a length of copper-clad steel antenna wire wound on a reel and passed out through a fair-lead installed in the bottom or side of the fuselage. The wire terminates in a wind sock, for horizontal polarization, or in a streamlined weight if vertical polarization is desired. The antenna is fed from a coaxial cable through a contact located where the wire enters or leaves the fair-lead, and may be considered as working against the skin of the ship as ground.

When trailing wires are operated at a fixed length less than $\lambda/4$ their impedance characteristics are marked by low resistance and by large capacitative reactance, varying rapidly with frequency. If transmitting efficiency is desired, the antenna must be fed through a matching section or tuning unit containing manually adjusted or motor-controlled variable lumped elements. When trailing wires are operated at some resonant length, such as $\lambda/4$ or $3\lambda/4$, they are, of course, nonreactive and have a reasonably large input resistance which varies with frequency in a complicated manner depending upon the size of the plane, the length of the wire, and the relative positions of plane and wire. Under many conditions this resistance is close enough to 50 ohms so that the antenna may be fed directly from the transmission line without recourse to matching sections. If extreme transmitting efficiency is required the antenna resistance may be matched to that of the feed line, by means of simple circuits of coils and capacitors, as is shown by the example of Figure 35.

At the low frequencies at which trailing-wire antennas are ordinarily used, matching sections consisting of lengths of transmission line are too bulky to be practical.

The patterns of simple trailing antennas in the lower v-h-f range are generally messy as compared with those of the fixed aircraft antennas recently developed for low-frequency use.

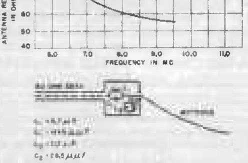


Figure 35. Characteristics of resonant trailing-wire antenna operated at $3\lambda/4$ resonance. (Data from ARL.)

Stingeree Antenna

The stingeree antenna, developed by the Bell Telephone Laboratories, is intended for broadband use, for either vertical or horizontal polarization, in the lower v-h-f range.

The antenna consists of a $\lambda/2$ dipole of the skin-back type, trailed from the plane at the end of 50 to 100 ft of standard coaxial cable. The antenna, sketched in Figure 36, contains a two-element transmission line matching section which is built into one side of the dipole. The radiating surfaces of the dipole consist of cylindrical metal-braid sheathing, quite similar to the armor used on RG-35/U coaxial cable. The far end of the antenna terminates in a streamlined weight. The combination feed and tow cable is coiled just before entering the dipole proper, the coil acting as a high-impedance choke in series with the antenna and

therefore tending to keep radiating currents from the outer surface of the feed line.

This antenna is said to have band widths of the order of 25 to 35 per cent at the extreme low end of the v-h-f band, the *SWR* at the input of the feed line, some 50 to 100 ft from the antenna, being less than 2:1 over such ranges.

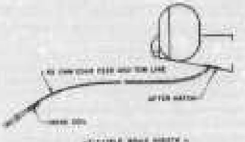


FIGURE 36. Stingeree antenna of Bell Telephone Laboratories.

The pattern of the stingeree is said to closely resemble that of a $\lambda/2$ dipole in free space. The antenna has the further advantage in that it is towed λ or more behind the ship and therefore its radiational characteristics may be expected to be much less dependent upon the nature and size of the plane than are those of ordinary fixed aircraft antennas.

14.9 SURFACE ANTENNAS

By mounting an aircraft antenna inside the plane, with its radiating surfaces flush with the skin of the ship, many of the problems of antenna design, including wind-drag, mechanical strength, icing, precipitation static, and tactical conspicuity, are solved at once, simply by elimination. These spectacular advantages have aroused great interest in surface antennas, an interest which has extended to the development of planes especially designed to accommodate such antennas, an example being the Bell D-6, a plywood plane upon whose nonconducting surfaces antennas were simply to be painted. Relatively little, however, had been accomplished in the field at the time the present report was prepared. In a rather complete file of the reports issued by the various laboratories engaged in aircraft antenna research there was not a single one dealing with surface antennas at frequencies lower than 3,000 mc.

The following material constitutes what little was learned about surface antennas at this laboratory (RCAL). It represents work done here largely at the request of the Radio Test Department, U. S. Naval Air Station, Patuxent River, Maryland.

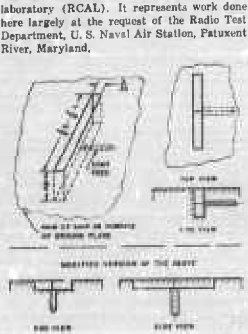


FIGURE 37. Single-slot antenna and modified version where $\lambda/4$ cavity has been bent back parallel to ground plane.

14.9.1 The Single-Slot Antenna

Figure 37 shows a sketch of a simple slot, approximately $7\lambda/8$ long by $\lambda/30$ wide, cut out of the skin of the ship. The slot is backed by a

rectangular resonating cavity, of the same cross section, and $\lambda/4$ deep. The system is fed by a short cylindrical radiator, running across the slot, and introduced along the center line of the wide side of the cavity as an extension to the inner conductor of the coaxial feed cable.

Figure 38 shows that the system behaves as an antiresonant circuit of fairly high Q.

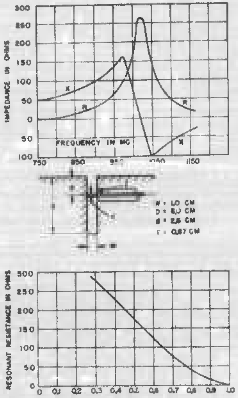


FIGURE 38. Impedance characteristics of simple-slot antenna.

This figure also shows the effect of the position of the feed, relative to the bottom of the cavity, upon the impedance level of the system. Since the input resistance at antiresonance decreases from high values to zero as the feed radiator approaches the bottom of the cavity, it is evident that the simple slot can be matched to the feed line, at one frequency, simply by adjusting the position of the feed.

Because of the steep characteristics of the antenna input impedance it is possible to obtain only four percent band width by means of a conventional $\lambda/4$ transformer, a band width which may be approximately doubled if a two-element transmission line matching section is used. In view of the size of the antenna

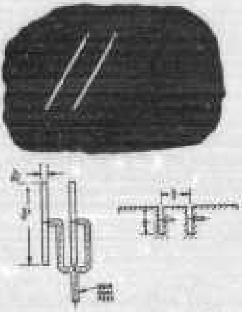


FIGURE 39. Double-slot antenna; two slots fed in phase.

in wavelengths, it is evident that a simple-slot antenna has impedance characteristics very much less suitable to broad-banding than those of conventional cylindrical antennas.[a]

[a] It may be objected that the above data on band width apply to an extremely unfavorable case, in that the impedance level of this slot is much higher than that of the line to which it is to be matched. It might appear that band width could be increased by changing the position of the feed point so that the resonant resistance more nearly approaches the characteristic impedance of the line. This is not the case; while the resistance match can be improved by lowering the feed point, the reactance is not affected in the same proportion; the result is that low-impedance antennas have *higher Q's* and *less* band width.

The radiation from slot antennas is confined to the same side of the ship as that upon which the antenna is located.

18.9.2 Double-Slot Antenna

Figure 39 shows a sketch of a system of two parallel slots, each $3\lambda/4$ long, spaced $\lambda/4$ apart, fed in phase. Although no impedance measurements have been made for this antenna, it is not likely that the band width of this system will be greater than that of a single slot, since there is no compensatory effect in the in-phase feeding of identical antennas, and since the effect of the mutual impedance at such small spacing is likely to be adverse.

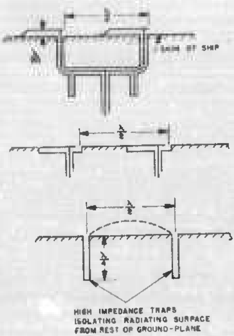


FIGURE 40. Designs of double-slot antennas by Lindenblad.

The field patterns of a double slot mounted under the wing of a PBY-5A are more symmetrical than those of the single slot, and the downward beam in the vertical plane transverse to the slots is sharper.

18.9.3 Lindenblad's Double-Slot Antenna

Figure 40 shows alternative arrangements of two slots spaced $\lambda/2$ apart, each slot being fed through a $\lambda/4$-deep resonant cavity which is folded back parallel to the skin of the ship. In this system there is evidence that surface currents in the $\lambda/2$-wide strip are responsible for most of the radiation, the strip behaving much like an array of thin $\lambda/2$ dipoles lined up side by side. The $\lambda/4$ feed cavities serve to isolate the radiating surface from the rest of the surface of the ground plane, performing the double function of placing a high impedance in series with the immediately adjacent outer surfaces and of insuring that what current does exist in these surfaces will be in phase with that in the strip.

SWR measurements indicate that band widths of the order of 10 to 15 per cent may be obtained without recourse to matching sections.

18.9.4 Lindenblad's Broad-Band Slot Antenna

A very interesting slot system, which includes a novel broad-band feed, is sketched in Figure 41.[8] From the outside of the ship the antenna appears as two thin slots, approximately 0.65λ long, spaced 0.15λ apart. From the interior of the ship the antenna appears as a thin square box, approximately 0.55λ on a side and 0.07λ thick, so oriented that the two outer slots lie parallel to one diagonal. This box is divided into two layers of approximately equal thickness by means of an inner sheet of metal, which contains an inner slot 0.06λ wide lying under the strip separating the two outer slots. A septum attached to this strip passes down through the inner slot to the bottom of the box. A feed strip, shaped as an equilateral triangle, leads from one edge of the inner slot to the bottom corner of the lower layer of the box, where it ties on to the inner conductor of a standard coaxial cable connector. By systematically varying the width, length, and spacing of the outer slots, the spacing of the inner slot, and the shape of the feed triangle, it has been possible to attain band widths of 20 per cent without need for external matching

sections. Much wider band widths are possible if the standard of matching were to be slightly relaxed, say to 2.5:1 *SWR* on a 50-ohm line.

The *SWR*-frequency characteristic of a broad-band slot system designed for altimeter use is included in Figure 41.

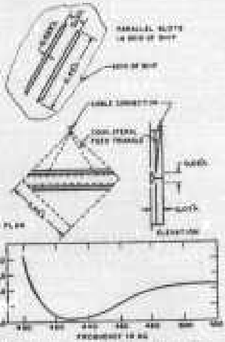


Figure 41. Plan and elevation views of broad-band slot antenna of [illegible].

18.9.5 Louvre, or Wedge, Antenna

The louvre antenna developed by P. S. Carter for an application quite remote from communications is an interesting example of a flush-mounted antenna. The system, Figure 42, consists of three very thin wedges arranged to overlap so that their open bases are spaced approximately $\lambda/4$ apart. The system is intended to be mounted upon the side or undersurface of the plane, depending upon the polarization and pattern desired, the open ends of the wedges appearing as long thin slots covered with low-loss dielectric, the antenna being fed directly from a coaxial line entering the middle wedge. The impedance characteristics of the antenna are such as to make it very sharply resonant. The field pattern, except in the plane normal to both the surface on which it is mounted and the length of the louvre openings, consists of fairly sharp single lobes, the positions of which in space may be adjusted simply by manipulating the tuning condensers in the two outer wedges.

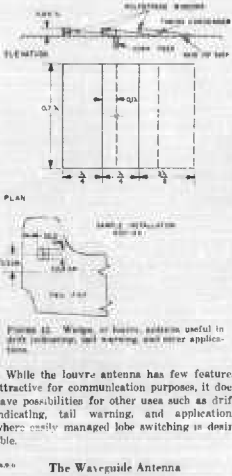
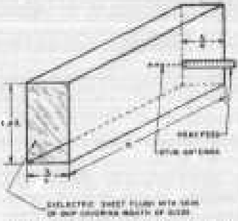


Figure 42. Wedge, or louvre, antenna useful in drift indicating, tail warning, and other applications.

While the louvre antenna has few features attractive for communication purposes, it does have possibilities for other uses such as drift indicating, tail warning, and applications where easily managed lobe switching is desirable.

18.9.6 The Waveguide Antenna

The waveguide antenna sketched in Figure 43 is a special type of horn antenna, i.e., a horn of zero flare. It is excited in the H_{01} mode by



[illegible] horn of [illegible]

means of a stub antenna mounted parallel to the short side of the guide, located approximately $\lambda/4$ from the closed end and fed directly from the coaxial feed line entering at the center of the long side.[a]

The open, or radiating, end of the guide can be covered with a sheet of low-loss dielectric material mounted flush with the skin of the ship. The patterns of a waveguide antenna having the dimensions shown in Figure 43 and mounted in the tail of an F6F are shown in Figure 44. This antenna was intended for horizontal polarization at 400 mc, the guide being oriented so that its long side is vertical. The patterns are quite similar to those predicted by the theory and experiment of Barrow and Greene.

18.7 Antennas in Semicylindrical Cavities

Figure 45 is a conventional cylindrical antenna mounted axially in a semicylindrical recess in the skin of the ship. The recess or cavity has an aperture approximately 0.4λ square which can be covered with a dielectric sheet mounted flush with the surface of the plane. The *SWR*-frequency characteristics show that while no band width is attainable with a simple stub radiator in the cavity (the resistance is too low and the reactance variation too steep in this case), the use of sleeve antennas results in quite appreciable band widths.

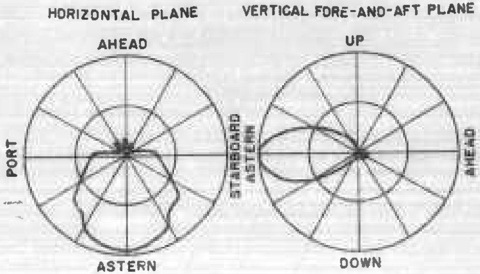
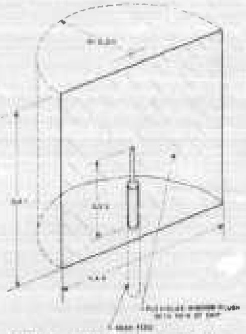


FIGURE 44. Antenna power pattern of waveguide antenna having dimensions of Figure 43, operating on 400 mc, located in F6F tail with mouth of guide facing directly aft.

[illegible] Conclusions

The material presented in the preceding sections summarizes what is known about surface antennas at this laboratory at the present time.



[illegible] cavity.

From this data we draw the following conclusions:

1. Surface antennas, whether they be slots, horns, or cavities, are much larger relative to the operating λ than are conventional exterior antennas. The maximum dimension is usually of the order of a $\lambda/2$ or more. While slots or horns of such aperture are feasible at frequencies down to 100 mc (assuming their use on *large* aircraft), a 30-mc slot antenna would require quite a little mechanical engineering. Surface antennas also have more or less bulk inside the skin of the ship, a fact which means that installation of even a small h-f antenna will be something of a major operation.

2. Surface antennas have much less intrinsic band width than ordinary antennas. For most communication purposes this is not much of an objection, particularly in the light of recent developments resulting in band widths of the order of 20 per cent or more.

3. Surface antennas have field patterns characterized by more directivity than is usually desirable in communication work. They do not transmit or receive energy in directions opposite to that in which they face, a situation which can probably be remedied by mounting two antennas on opposite sides of the ship.

4. Surface antennas, while having reached a stage of development permitting their immediate application to many aircraft antenna problems, constitute a rich and virgin field of research, particularly along the lines of increasing band width (by continued development of broad-band methods of feeding them), reducing bulk (possibly by means of filling them with low-loss dielectrics of high inductive capacity), and improving patterns by means of multiple-antenna systems.

[illegible] POWER CAPACITY OF AIRCRAFT ANTENNAS

The maximum power that can be handled by aircraft antennas depends upon the nature of the antenna and upon atmospheric conditions.

Power capacity varies approximately as the square of the conductor diameter, and consequently will be greater for thick cylindrical and conical antennas than for antennas consisting of one or more small wires, such as fixed- or trailing-wire antennas or fans.

Since breakdown due to corona or arc-over depends upon field strength rather than voltage, maximum power will depend upon the orientation of the antenna with respect to the ground plane against which it is worked, being greater for simple vertical antennas than for antennas having components parallel to the skin of the ship. Furthermore, since antenna voltage for a given power input is a function of the current distribution along the antenna, it is evident that an antenna with top-loading will have a different power limit from that of a simple $\lambda/4$ stub.

18.10.1 Antenna Length and Resistance

The antenna voltage for a given power input is proportional to the square root of the input resistance, implying that the maximum power for a given corona voltage will be proportional to the radiation resistance of the antenna. Hence a $\lambda/4$ or longer antenna will handle more power than a short antenna. Since an electrically short antenna requires inductive loading to be fed at all, and since the ohmic resistance of the coil may be of the same order of magnitude as the radiation resistance of the antenna, an appreciable fraction of the input power will not reach a short antenna at all. The loading coil must be designed to dissipate that fraction safely.

18.10.2 Antenna Surface

The surface of a high-power antenna should be smooth and of relatively larger radius of curvature, since corona sets in at lower voltages the rougher the surface.

The effect of dirt on the antenna surface is to start local discharges and may cause the onset of general corona at lower voltages.

18.10.3 Atmospheric Conditions

The breakdown voltage of air is a complicated function of its density, and so depends upon pressure and temperature. The dielectric strength of air increases with density, density decreases with decreasing pressure and with increasing temperature. The power capacity of aircraft antennas is therefore less at high altitudes than at low, the effect of decreasing pressure much more than compensating the effect of decreasing temperature as the altitude increases.

While moisture present in the air has little effect upon the starting point of corona, once corona is started rain and humidity reduce the spark-over voltage greatly.

Ionization pre-existing in the air surrounding the antenna has little effect on the onset of corona. However if the plane picks up sufficient charge, corona in the form of precipitation static will set in, regardless of the voltage on the antenna.

18.10.4 Summary

The problem of power limits of aircraft antennas is too complex to permit solution by simple, unqualified rules or formulas. It is, however, possible to make simplifying assumptions which may be useful in a qualitative way in showing the effect of a few of the factors entering into the problem. The curves of Figure 46 represent such approximations. While in particular cases they may not even be correct as to order of magnitude, they show, in a general way, the relation between power limit and antenna diameter for two types of antennas important in aircraft radio.

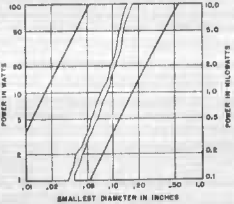


FIGURE 46. Power-handling limit of $\lambda/4$ antenna at elevation of 40,000 ft.

18.11 PRECIPITATION STATIC

Precipitation static interferes with aircraft communication when the receiving plane passes through rain, snow, or through clouds of dust or ice particles. When first observed the static appears as a series of popping noises in the receiver, which noises finally develop, as flight continues, into a continuous roar completely obscuring the signal. The effect seems to be

worse with higher speed planes, and in a given case static can usually be reduced by reducing speed.

16.12.1 Remedies

The elimination of precipitation static is achieved by a twofold attack on the problem.

1. Sharp points on the surface of the plane are removed. As far as antenna design is concerned this demands that the use of fine wire or of fittings involving surfaces of small radius of curvature be avoided.

2. Provision is made for dissipation of the charge accumulated on the plane in a noise-free manner at a point remote from the antennas. Such discharge can be effected by means of a very thin wire, ending in a sharper point than any on the surface of the plane, trailing from the rear of the plane. A large-value resistor in series with this wire tends to damp the oscillations ordinarily associated with the discharge.

16.13 AIRCRAFT ANTENNAS AND AIR DRAG

All antennas projecting beyond the surface of the airplane are aerodynamic liabilities in that they are sources of parasitic drag. At ordinary subsonic velocities parasitic drag may be considered as consisting of two distinct types; frictional drag and form drag, which although interrelated in their effects will be considered separately.

16.13.1 Frictional Drag

Frictional drag is the resistance experienced by a moving body due to the viscosity of the air through which it moves. It is always proportional to the total surface area exposed to the airstream. Any moving surface is surrounded by a transition layer in which the air velocity relative to the surface increases from zero at the surface (neglecting the phenomenon of slip) to the full value of the stream velocity at the outer edge of the boundary layer. For low Reynolds numbers (the product of the air density, the stream velocity, the maximum linear dimension of the body normal to the stream, and the reciprocal of the coefficient of viscosity of the air) the flow in this boundary layer is laminar, consisting of layers in which all or almost all the fluid motion is parallel to that of the stream. Under this condition the coefficient of frictional drag is almost independent of the nature of the surface of the body, depending only upon the Reynolds number and the shape of the body. At higher Reynolds numbers, above a certain critical velocity which depends upon the shape of the body, the flow in the boundary layer becomes turbulent and there is greater frictional drag. For turbulent flow, frictional drag is greater the rougher the surface.

16.13.2 Form Drag

Form drag is due to the disturbance created in the airstream by passage of the moving body and depends largely upon the shape of that body. For objects with sharp edges the form drag is virtually independent of Reynolds number, being almost entirely due to the difference in pressure upon the leading and trailing surfaces. For rounded bodies the form drag coefficient depends upon the Reynolds number, the surface roughness, and the degree of turbulence in the airstream. Such rounded bodies as spheres and cylinders may have smaller drag coefficients at high velocities than at low, the reason being that at low Reynolds numbers the boundary-layer flow is laminar, the flow separating on the leading side of the body, resulting in a wide wake and a large form drag, while at higher Reynolds numbers the boundary flow is turbulent and does not separate until it reaches the trailing side of the body, resulting in a narrow wake and a correspondingly smaller drag. The magnitude of this effect can be startling. In the case of a sphere the drag coefficient suddenly decreases sixfold when the velocity reaches the critical value at which turbulence sets in. Turbulence pre-existing in the airstream reduces the critical velocity at which this decrease in form drag occurs. At still higher velocities, beyond the critical velocity, the drag coefficient rises slowly with increasing

velocity until sonic speeds (75 per cent or more of that of sound) are reached.

At sonic velocities the entire character of the airflow around the moving body changes, the leading surfaces setting up shock or compression waves resulting in a type of drag known as wave drag. The wave-drag coefficient rises rapidly as the velocity of the body approaches that of sound, the total drag becoming much greater than that ascribable to form or friction.

18.18.3 Calculation of Antenna Drag

Tables of values of frictional- and form-drag coefficients are available in the literature of aerodynamics which also includes formulas by means of which the drag on a given antenna may be computed with reasonable accuracy for either laminar or turbulent flow. Wave drag is a relatively new phenomenon, encountered since the start of the war with the attainment of sonic velocities in dives with super-fast planes. Very little quantitative data information on wave drag was available in published literature at the time this report was prepared.

It should be remarked here that the total drag of a body moving at ordinary speeds may consist of frictional drag and form drag in almost any proportion, ranging from 100 per cent frictional drag for a properly designed streamline form to almost 100 per cent form drag for a smooth sharp-edged plate.

An approximate semi-empirical formula for the total drag of a smooth circular cylinder—a shape common among conventional aircraft antennas—is

$$D = 0.0030V^2$$

where D is the drag in pounds per square foot of projected area and V is the velocity in mph. This formula results in good agreement with experiment for wire and rod antennas moving at moderate speeds.

18.18.4 Measurement of Antenna Drag

A more direct procedure, giving more satisfactory results when the antenna is of complicated shape or located in such a position on the plane that the assumptions underlying the drag formulas are not fulfilled, is to measure the drag of a model of the antenna mounted on a scale model of the plane by means of a wind tunnel. An alternative method is to mount the antenna on the actual plane and put it through all the maneuvers likely to be met in ordinary flight.

18.18.5 Simple Means of Reducing Drag

Frictional drag may be reduced by smoothing the antenna surface. While at low speeds the nature of the surface is more or less immaterial, at high speeds (turbulent flow) a smooth surface is essential to low drag.

Form drag may be greatly reduced by streamlining, that is, by so shaping the antenna that it produces little eddy-current disturbance as it passes through the air. The shape of the streamline form resulting in minimum drag depends upon the velocity, a ratio of major-to-minor axis of 2 or 3 being satisfactory for moderate speeds, of the order of 200 mph; larger ratios, i.e., thinner forms, are required at higher speeds.

The effectiveness of streamlining in reducing drag is evident in Table 1, in which the drag in pounds per projected foot for standard circular aircraft cable is compared with that for streamlined wire of similar nominal diameters.

TABLE 1. Effect of streamlining on antenna drag

Nominal diameter in inches	Drag in pounds per projected foot at 100 mph	
	Circular cable	Streamlined cable*
0.25	0.64	0.056
0.3125	0.80	0.067
0.375	0.96	0.077
0.50	1.21	0.092

* Axis ratio 4:1.

While antenna wires are rarely streamlined in practice because of the difficulty of maintaining the wire orientation in flight and because of the fact that the drag of the antenna fittings are usually much greater than that of the wire itself, it is worthwhile to streamline antennas of thick cylindrical form.

12.12.6 Other Factors Affecting Antenna Drag

Obviously the length, cross section, and orientation of the antenna are important factors in determining its total wind resistance. Unfortunately these parameters are determined by electrical considerations, which—since they are inextricably connected with the reason for having the antenna on the airplane in the first place—must be regarded as being at least as important as the matter of air drag.

Antenna length is controlled by frequency. A conventional transmitting antenna, to be reasonably efficient and capable of even modest band width, must be of the order of $\lambda/4$ long. Since wavelength is inversely proportional to frequency, the drag of a smooth-surfaced cylindrical antenna will also be inversely proportional to frequency, other factors remaining constant. This situation is emphasized in Table 2 in which the approximate drag of smooth cylindrical antennas 1 in. in diameter, moving at 200 mph, are compared for various resonant frequencies.

TABLE 2. Antenna drag as a function of frequency. Drag on vertical quarter-wave antennas 1 in. in diameter at 200 mph.

Frequency (mc)	Antenna length (m)	Drag (lb)	Wasted power (hp)
3,000	1	0.831	0.44
300	10	8.3	4.4
30	100	83.0	41.0

These figures are only approximate, and in certain antenna installations may not even be of the right order of magnitude. They are intended only to support the following rule: *If drag must be minimized, avoid low frequencies* particularly if the impedance and pattern requirements are such as to demand the use of a conventional exterior antenna.

The cross-sectional dimensions of conventional antennas are controlled by the band width desired. Neglecting special cases in which the antenna impedance is markedly affected by its location on the particular plane involved, the fatter the antenna the greater its intrinsic band width, assuming that it can be fed in a manner which will not detract from that intrinsic band width. Other things being equal, if the impedance characteristics of a given h-f antenna of cylindrical form are to be duplicated at a lower frequency, the cross-sectional dimensions of the antenna must be scaled up in the same proportion as its length. The consequences of this fact on air drag as a function of frequency, for constant band width, is shown in Table 3.

TABLE 3. Antenna drag as a function of frequency. Drag on vertical quarter-wave antennas of comparable band width at 200 mph.

Frequency (mc)	Antenna length (m)	Antenna diameter (in.)	Drag (lb)	Wasted power (hp)
3,000	1	0.25	0.21	0.11
300	10	2.5	21	11
30	100	25.0	2,100	1,106

Again the purpose of the table is purely illustrative, to show that, for conventional antennas at least, *the desires for band width at low frequencies and for low drag are incompatible.*

Antenna orientation is controlled by the nature of the pattern and by the type of polarization required. For vertical polarization in the v-h-f and u-h-f bands, orientation is usually not a factor, since a vertical $\lambda/4$ antenna is necessarily perpendicular to the airstream. For vertical polarization in the lower v-h-f and higher h-f ranges, where flat-top antennas must be used, the horizontal portion of the antenna should be strung back parallel to the line of flight. For horizontal polarization the antenna should be oriented so that it presents a minimum projected area to the wind stream, providing such orientation is consistent with the nature of the field pattern desired.

12.12.7 Special Low-Drag Antennas

In many instances it is possible to satisfy the polarization, pattern, and impedance requirements of a given problem by means of antennas having much less air drag than the simple wires, stubs, and whips to which the preceding discussion applies. A classic example of a satis-

factory low-drag antenna is the wire fan developed by the Antenna Section, Research Division, Aircraft Radio Laboratory, and described elsewhere in this report. Other antennas having reduced drag to the extent that they are shorter than conventional radiators, are treated in Section 18.5.3.

The most effective way to minimize drag is to remove the antenna from the airstream, placing it inside the skin of the ship. Several such antennas are described in Section 18.9.

In many cases low-drag antennas of the types mentioned above will have impedance, pattern, or mechanical features making them unsuitable for a particular application, in which cases about all that can be done is to reduce friction by smoothing the antenna surface and to reduce form drag by streamlining.

Chapter 19

DEVELOPMENT OF FAIRED-IN ANTENNAS

Development of a suitable device for exciting the surface of an all-metal plane to serve as the radiator. Slots, bars, etc., are looked upon as exciting devices and not as the primary radiators. Field pattern, surface current, and impedance measurements were made on scale-down models at a wavelength of 10 cm and on full-scale plane models using v-h-f frequencies.

19.1 INTRODUCTION

If antennas in the v-h-f band are to be used on all-metal high-speed aircraft it is necessary that the antennas be streamlined into the contour of the airplane. This means that the surface of the plane becomes an important component of the radiating system. In fact, the approach in this project[a] was to consider the current on the surface of the plane as the principal source of radiation. The plan was to investigate the possibility of designing an exciting device that would not protrude, that would be compact, that would have suitable impedance characteristics, and that would give the required field pattern.

Although the work was only well under way at the close of the war and the project terminated, much work was accomplished that will serve as background for its continuance under the Office of Naval Research.

19.2 DEVICES INVESTIGATED

The following current-exciting devices were investigated.

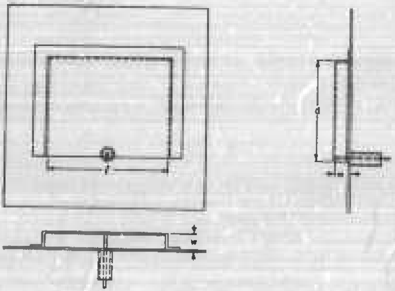


Figure 1. Schematic drawing of W-slot antenna.

19.2.1 On Sheets at 10 Cm

Slot Antennas

Slot antennas are recognized by the presence of these features: (1) the surface to be excited, (2) a cavity, (3) a slot to couple the cavity to the surface, and (4) a dipole or other exciting device to set up a field in the cavity.

[a] Project 13-110, Problem No. 5, Contract OEMsr-1441, Harvard University.

W Slot. The advantage of this type of slot over most others is that the skin of the airplane need not be cut except for the coaxial line. Two W slots were investigated having the following dimensions (see Figure 1):

W-1 (short slot)
$l = 0.10\lambda \quad d = 0.29\lambda \quad w = 0.03\lambda.$

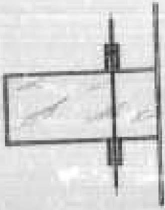
W-2 (long slot)
$l = 0.60\lambda \quad d = 0.45\lambda \quad w = 0.08\lambda.$

A Slot. The essential dimensions of this type of slot are shown in Figure 2. Two models were examined having these dimensions:

A-1 (narrow slot) $w = 0.02\lambda \quad l = 0.64\lambda.$
A-2 (broad slot) $w = 0.315\lambda \quad l = 0.64\lambda.$

reversing sleeves, one sleeve located at each end of the bar beneath the surface. The purpose of the phase-reversing sections is to increase the radiation resistance by spreading the current on the surface. If these sleeves are not present, the surface current is merely the image current of the bar. This current falls off very rapidly as one moves away from beneath the bar. The radiation resistance is low without the sleeve because the image current and the bar current are in opposite directions and for this reason the field of the bar tends to be cancelled by the field of the image current.

The following bar antennas were tested.



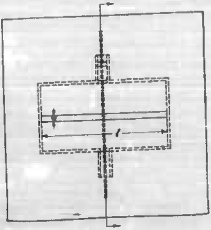


Figure 2. Schematic drawing of A-slot antenna.

H Slot. This type of slot (Figure 3) was used in an investigation of the proper location of an AN/APN-1 altimeter in a P4M bomber. (See Chapter 20.)

C Slot. This slot (Figure 4) was devised to see what effect bending the slot back on itself would have on the surface current and field pattern.

Bar Antennas

A simple bar antenna consists of a metal rod parallel and very close to the surface to be excited, together with two coaxial type phase-

1. B-1 end-fed bar. This was adapted from the B-2 center-fed bar by using an unbroken bar and by covering up the center section. The coaxial cable from the transmitter was connected to one of the coaxial fittings, the other fitting was connected to a variable-length line. This latter connection permitted locating the maximum current at the center of the bar. The bar length was about 0.5λ.

2. B-2 center-fed bar. The coaxial cable from the transmitter was connected to the center as shown in Figure 5. The two side fittings may be connected together by an adjustable length of coaxial line or each fitting may be attached separately to an adjustable length of line. In

either case the adjustment is for optimum location of the current maximum.

3. 2B-1 parallel-bar antennas. Two B-1 bars separated by about 0.5λ (adjustable), driven phase difference adjusted for circular polarization.

5. ½B-1 or half-bar. For proper adjustment of the B-1 antennas a voltage node appears in

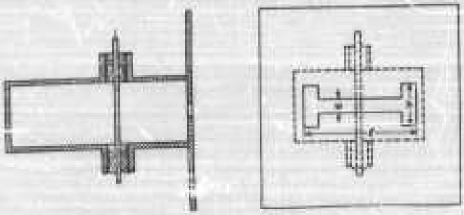


FIGURE 3. Schematic drawing of B-1 antenna.

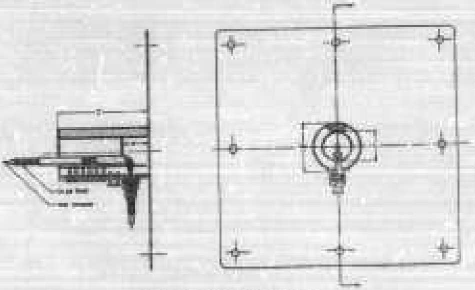


FIGURE 4. Drawing of [illegible] antenna.

either in phase or 180° out of phase. The length of the bars was about 0.5λ.

4. 2BX-1 crossed-bar antennas. Two perpendicular B-1 antennas of standard size with the middle of the bar. Therefore it can be connected directly to the surface at this point. This makes possible an antenna of half the length of the B-1 bar.

19.2.2 On Sheets in the V-H-F Band

B-1 and B-2 antennas were tested for their proper impedance characteristics. Various sizes and shapes of bars and phase-reversing sleeves were tested.

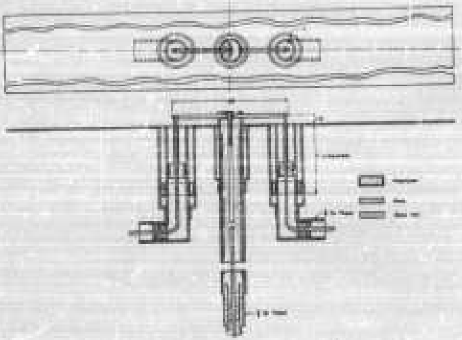


FIGURE 5. Drawing of center-fed B-2 bar antenna.

19.2.3 On a Scaled-Down Model of a P47N.

SLOT ANTENNAS

A W slot was adapted to the vertical stabilizer and tested for 360° coverage with good results.

BAR ANTENNAS

The B-1 end-fed bar was mounted vertically on the sides of the fuselage in the following positions:

Position 1 front left 0.23λ from the nose of plane.
Position 2 front left 0.87λ from the nose of plane.
Position 3 front right 0.47λ from the nose of plane.
Position 4 rear right 2.75λ from the nose of plane.

Total length of plane was 5.3λ. Combinations of the above positions using [illegible] B-1 antennas were also used.

STUB ANTENNA

A vertical stub λ/4 in length was mounted on top the fuselage at 2.87λ from the nose. This antenna was useful for comparison purposes.

19.3 METHODS OF MEASUREMENT

Measurements undertaken under the project were (1) the field patterns of the antennas as well as the effect of different conditions of adjustment, (2) the distribution of surface cur-

rent on the sheet of metal or the model plane on which the antenna was mounted. The results of such measurements were to be properly correlated.

10.5.1 Field Pattern Measurements

The antenna under test was excited through a 50-ohm coaxial cable from a modulated klystron oscillator. A double stub tuner matched the line to the oscillator. Part of the oscillator output was fed to a crystal detector circuit which was part of the modulation monitor unit. The relative output of the oscillator could be checked by the detector current as read on a meter or by the deflection on the screen of a cathode-ray oscilloscope.

The operating wavelength of the klystron oscillator was 10 cm while the square wave modulation could be varied in frequency from 500 to 5,000 cycles.

The antenna system was mounted on a remotely controlled turntable driven by geared-down synchronous motors and equipped with selsyns for remote indication of the azimuth position of the antenna.

In addition the antenna could be rotated about a horizontal axis to give an "elevation" rotation.

The receiving antenna consisted of a half-wave dipole at the focus of a 120-cm parabolic reflector and placed at a distance of about 130λ from the transmitter to insure far-zone conditions. The received signal was detected by a bolometer consisting of a 10-ma Littlefuse inserted in the coaxial fittings of the paraboloid. The output of the bolometer was amplified (Figure 6) and the amplifier output was read on a d-c milliammeter or was recorded on an

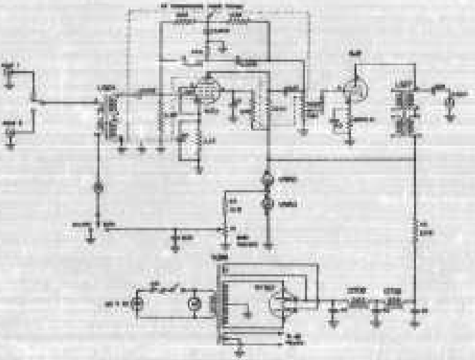


FIGURE 6. Amplifier for use with bolometer.

external automatic recorder. The readings were directly proportional to the square of the electric field.

SURFACE-CURRENT MEASUREMENTS

In securing surface current measurements it is of primary importance that the indicating instrument response be determined by the currents flowing in the small area of the surface surrounding the point investigated. A small loop placed in a plane perpendicular to the surface being excited satisfied this requirement. At S-band frequencies, a loop area of from 1 to 2 sq cm gave good localization of response with sufficient sensitivity to permit measurement of very weak currents.

The detecting element was a crystal. The square-wave modulation of the klystron (type 410R at 80-100 watts input) made it possible to amplify the detector output.

The final report gives the results of a great many tests and includes surface current diagrams.

19.3 Impedance Measurements

Delays in procuring a P47 plane limited the impedance measurements planned to actual v-h-f band data on B-1 and B-2 antennas on flat sheets. Since measurements on a scaled-down model of the antennas at S-band frequencies were seen to involve many difficulties, investigations along these lines were undertaken only as a sideline. The results of the measurements made are contained in the final report[1] which also contains the theoretical analyses necessary for proper evaluation of the work accomplished.

19.4 CONCLUSIONS

Although the planned work was not completed, due to the termination of the contract at the end of the war, certain conclusions were reached. One of the most striking results of the surface current investigation is that, at v-h-f frequencies, skin antennas usually excite the plane as a whole. Thus the behavior of such antennas is dependent on the general shape of the aircraft. A single antenna located properly on the side of the fuselage may be able to so excite the plane itself as to give 360° coverage with no serious nulls. Also there may be a large horizontally polarized field from an antenna that, judging from the result on a flat sheet, should give only vertical polarization.

On the other hand the interpretation of the surface-current patterns given in the final report[1] is seriously limited by a lack of knowledge of phase relationships. Further research on surface currents should include the measurement of relative phase along with measurement of direction and magnitude of current.

Slot-antenna studies were very largely conventional. The bar antennas, however, represent a significant departure from other skin antennas. Their importance lies not only in their merits as possible antennas but also in the fact that they emphasize the significance of current on the plane itself. With this viewpoint in mind, many other novel ways of exciting a plane at v-h-f should result from further research. Such research should lead to antennas capable of satisfying a great variety of requirements.

The B-1 and B-2 bar antennas differ decidedly in their behavior from that of a dipole mounted at the same distance (0.02λ) from a flat sheet. The difference expresses itself in a greater extent of surface current excitation, a narrower beam width, and greater efficiency.

Impedance measurements of the bar antennas were made only in the v-h-f band and at full scale where the phase-reversing sleeves were very much smaller in diameter relative to the wavelength and the distance of the bar from the sheet than they were in the 10-cm models. Here the band width for a 2:1 standing wave ratio was about 0.8 per cent. Thus the bar with phase reversing sleeves of very small diameter is serviceable as it stands at one frequency only. If the reversing stubs are tuned by remote control, a very broad band is possible.

The fact that a plane may be excited as a whole by a bar antenna or other device makes it seem likely that its impedance is changed

considerably from that as measured on a flat sheet. To investigate this effect a full-scale plane mounted on a platform so that it is sufficiently decoupled from its ground image, or a careful scaling down of the plane, the exciting antenna, and the antenna feed systems, is necessary.

One of the conclusions that was drawn from the measurement of surface currents on the model plane was not only that currents may be large over the entire plane, but that currents remote from the antenna may be of primary importance in determining the field. Hence the shape of the plane may, at v-h-f frequencies, materially affect the impedance at the terminals of the exciting antenna.

In the appendices of the final report[1] will be found certain analytical studies useful to any work in this field. These subjects include "Surface current distributions that produce circular horizontal polarization"; "Broad-band characteristics of a dipole using a series transformer as a matching section"; "On the proper spacing of insulating beads"; and "A conversion chart for impedance measurement using transmission line."

Chapter 20

MISCELLANEOUS ANTENNA RESEARCH

20.1 LOCATION OF ANTENNA FOR AN/APN-1 ALTIMETER ON NAVAL AIRCRAFT

This project was set up to study the problem of locating H-type slot antennas for minimum feed-through on Navy type P4M aircraft.[a] The study indicated that the slots should be located on the opposite sides of the horizontal stabilizer undersurface and that for absolute minimum feed-through the slots would have to be approximately perpendicular to each other, thus seriously affecting the operation of the altimeter.

20.1.1 Introduction

Since the AN/APN-1 altimeter operates in the region 420 to 460 mc, it was decided to use a 1/7 scale-down in constructing a model P4M jet-assist bomber tail section and to employ a frequency of 3,080 mc. The plan was to measure the surface current distribution on the horizontal stabilizer surface and on the surface of the fuselage in the vicinity of the stabilizer for a wide variety of positions of the transmitter H slot. These measurements were to determine lines of flow and contours of constant surface-current amplitude. Then for each position of the transmitting slot, positions for the receiving slot would be chosen:

1. Only in regions permitted by the internal structure of the plane.
2. So that the angle formed by the lines of orientation of the transmitting and receiving slots would not exceed 45°.
3. At positions of minimum surface-current amplitude.
4. So that the receiving slot would be oriented parallel to the lines of surface-current flow in its vicinity.

[a] Project 13-110, Problem No. 7, Contract OEMsr-1441, Harvard University. Originally Project 13-108.

Conditions (1) and (2) constitute limitations imposed by the practical location of the slots and acceptable altimeter performance. Conditions (3) and (4) constitute limitations on the location of slots for minimum feed-through caused by surface-current coupling.

20.1.2 Laboratory Technique

To obtain accurate measurements of feed-through it was necessary to minimize direct reflection of energy from one antenna to the other. This made it desirable to simulate flight conditions by mounting the model on a platform far from ground and upside down so that the H slots would be directed skyward. Difficulties in getting the platform delayed this part of the study until near the time the experimental work was terminated.

Concurrent with surface-current measurements, attempts were made to determine a satisfactory method of measuring absolute feed-through. The problems of establishing a proper reference level, of matching, and of cable losses all had to be solved before the actual feed-through data could be collected.

As a first step in this direction a flat metal sheet was constructed so that it could be driven by a waveguide slot. Holes were cut in the sheet at positions of different current amplitude. A rotatable H-slot mounting disk was designed and constructed so that when located in any one of the holes it could be rotated continuously through 360° and clamped in any desired position. With this setup an investigation was made as to the correlation between feed-through and (1) surface-current amplitude at the receiving H-slot position and (2) the angle between H-slot direction and lines of surface-current flow in the vicinity of the slot. The results obtained are of value in estimating the extent to which the feed-through is minimized by locating a receiving H slot according to conditions (3) and (4) above.

20.1.8 Conclusions

The final report[1] gives a description of the experimental equipment employed and the results of the measurements to the end of the contract. Work of a somewhat more general nature is continuing under a contract with the Office of Naval Research.

The conclusions cited below are tentative.

Surface-current coupling between transmitting and receiving antennas can be minimized by choosing the slot positions in such a way that the receiving slot is located in a region of minimum surface-current amplitude and orienting the receiving slot so that it is parallel to the lines of surface-current flow in its vicinity.

Measurements made indicate that minimum feed-through values of between 70 and 80 db down may be reasonable and that values between 90 and 100 db down may not be beyond the realm of possibility.

It may be necessary to orient the slots at an angle with respect to the line of flight. Surface currents are smaller on the stabilizer surface opposite to the side on which the transmitting slot is located and less than on the bottom surface of the fuselage section adjacent to the horizontal stabilizer.

To achieve best final results in locating the slots, the measurement technique may have to be carried out on a full-scale mock-up of the tail assembly mounted at a sufficient height above ground so that the presence of the observer and observing equipment may be lessened in its power to affect the measurements.

20.2 STUDY OF PROBLEMS ARISING FROM CLOSELY GROUPED ANTENNAS

20.2.1 Introduction

Experience in the theaters has indicated that the first practical step in minimizing the severity of local radio interference in a headquarters area is to establish separate sites for groups of transmitting and receiving antennas. The principal purpose of the survey conducted under Project 13-103[b] was to determine the minimum required separation between such transmitting and antenna "parks," and between individual antennas within an antenna park. This separation is largely a function of certain spurious interference-producing properties of existing military radio sets, and on the coupling between various antenna types over different types of soil. Some data on apurious radiations and responses of radio sets were obtained in earlier work under Project C-79[c] (Contract OEMsr-1018), and these were supplemented by additional measurements on a number of sets. Because of the theater needs, this information on set characteristics was incorporated as part of War Department publication TM 11-486[c] prepared by the contractor prior to publication of the final report on Project 13-103.

[b] Project 13-103, Contract No. OEMsr-1412, Western Electric Co.

20.2.2 Results of the Survey

The final report[2] on the project contains information for estimating the required separation between transmitting and receiving antenna parks for both h-f and v-h-f tactical radio circuits, separations which should exist between individual antennas in an antenna park, and the relative advantages of the several methods of connecting several receivers to a single antenna.

Considerable information is given on transmitter-to-receiver interference as a result of spurious radiation at harmonics of the master oscillator, spurious outputs caused by interference between transmitters, effect of radiation from receivers, and spurious responses of superheterodyne types of receivers, with curves and charts enabling one to predict where such undesired receiver responses will occur in frequency.

Separation between transmitting and receiving antennas is considered from several angles and data given in tables and charts taking into account the types of antennas employed, the ground characteristics, the weakest usable signals, and the tolerable r-f interference-to-signal ratios.

Suggested layouts of h-f sky-wave transmitting or receiving antenna parks are given based on (1) assigned frequencies being divided into groups in such a way that the frequencies of

any pair within the group are not less than 10 per cent apart in frequency, (2) half-wave antennas for the frequencies within a group being placed parallel to each other and about 5 ft apart, and (3) antenna groups being located about 250 ft apart.

Similar layouts are given in the final report when other types of transmission are utilized, ground wave signals, for example, for other types of antennas, etc.

Considerable data are given on the mutual impedance between coupled antennas over an imperfectly conducting earth, and on possible methods of connecting several receiving antennas to a common antenna.

20.3 STUDY OF IMPROVISED V-H-F ANTENNAS

20.3.1 The Problem

Reports from combat areas indicated that dipole antennas made from ordinary field wire and using paired field wire for feed lines might be used when standard antennas and feed lines were not available. The purpose of this project[c] was to evaluate the losses in such systems and to suggest effective arrangements which could readily be improvised from available materials.

20.3.2 The Solution

The type of wire in common theater use consisted largely of ordinary field wire (W-110-B), long-range tactical wire (W-143) and spiral-four cable (WC-548). Measurements indicated that the losses in these wires would attain values as high as 10 to 25 db or more per 100 ft at 100 mc. At 30 mc, ordinary field wire when wet had losses as high as 15 db per 100 ft.

The high losses of ordinary wire used as a transmission line indicated the use of spaced leads for the feed line. An improvised antenna consisting of a half-wave dipole and a spaced line with a quarter-wave matching section at each end operated satisfactorily. The line consisted of two conductors formed by field wire spaced about 2 in. apart while the quarter-wave matching sections at each end of the line consisted of suitable lengths of paired W-110-B wire. With such a line 100 ft long, the power radiated from a transmitter was only a few db less than it would have been with a flexible coaxial cable. The actual losses were 2 to 4 db greater at 30 to 40 mc and 3 to 6 db greater at 70 to 100 mc. With the line wet these values were increased an additional 1 to 3 db. These losses were relatively unimportant when receiving unless the signals were marginal.

Loss characteristics, figures illustrating antennas hung from trees, etc., will be found in the contractor's final report.[c]

20.4 DISGUISED ANTENNAS

20.4.1 Introduction

The problem[d] was to design an antenna that would not project into the air, revealing the presence of the radio set to which it was connected. The research was confined to the portable radio set SCR-300 which is ordinarily used with one of two antennas, one being 10 ft 8 in. long, the other being 33 in. long with a parallel loading circuit grounded to the case of the set.

The tests were made largely in the field, one pack set using the improvised antenna and the other the standard 33-in. antenna. The testing procedure consisted in comparing the improvised antenna with the standard collector under the same conditions.

20.4.2 Results of Field Tests

The most promising disguised antenna tested was that employing helmet and counterpoise. A short length of wire connected the helmet to the parallel matching section. Another wire, connected to the ground terminal of the pack set and serving as counterpoise, extended almost to the ground. The latter could probably

[c] Project 13-102, Contract OEMsr-1411, Western Electric Co.

[d] Project 13-110, Problem No. 10, Contract OEMsr-1441, Harvard University.

be sewed into the trouser leg. Good signal strength and intelligibility was possible over ranges of 1 to 3 miles. If the operator's head was less than 15 in. above the ground the maximum range was about one mile.

Another promising arrangement was to use the pack set itself as antenna and to drive it against the ground or the operator's body. An L-type matching section was required. The range was about the same as that described above.

Tests in which the antenna wire was sewed into the clothing were not so successful as the other schemes devised.

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SERVICE PROJECT NUMBERS

The projects listed below were transmitted to the Executive Secretary, NDRC, from the War or Navy Department through either the War Department Liaison Officer for NDRC or the Office of Research and Inventions (formerly the Coordinator of Research and Development), Navy Department.

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A summary technical report of the National Defense Committee (NDRC) activities is presented. The methods and results are given which cover widely varied direction finder and antenna research and development. One part deals with the basic studies in direction finding, including means of measuring ground constants, and of rating DF systems in terms of wanted-to unwanted pickups. Also discussed is the effects of connecting cables with Adcock systems. Another section deals with physical equipment and systems developed under the direction of Division 13. Finally, the early work of sferics, the use of radio direction finding for locating storms, is reviewed.

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